NASA CONTRACTOR REPORT

NASA CR-141394

BISTATIC RADAR
SEA STATE MONITORING
FIELD TEST

(NASA-CR-141394) BISTATIC RADAR SEA STATE MONITORING FIELD TEST (Battelle Columbus Labs., Ohio.) 59 p HC $4.25 CSCL 08C

N75-23919

Unclas
G3/35 18530

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Prepared under contract No. NAS6-2006 by:

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for:

NATIONAL AERONAUTICS AND SPACE ADMINISTRATION
WALLOPS FLIGHT CENTER
WALLOPS ISLAND, VIRGINIA 23337

March 1975

1. Report No.: NASA CR-141394

2. Government Accession No.:

3. Recipient's Catalog No.:

4. Title and Subtitle: BISTATIC RADAR SEA STATE MONITORING - FIELD TEST

5. Report Date: June 1974

6. Performing Organization Code:

7. Author(s): G. T. Ruck, C. K. Krichbaum, and J. O. Everly

8. Performing Organization Report No.:

9. Performing Organization Name and Address:
Battelle
Columbus Laboratories
505 King Avenue
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10. Work Unit No.:

11. Contract or Grant No.: NAS6-2006

12. Sponsoring Agency Name and Address:
National Aeronautics & Space Administration
Wallops Flight Center
Wallops Island, Virginia 23337

13. Type of Report and Period Covered: Contractor Report

14. Sponsoring Agency Code:

15. Supplementary Notes:
In accordance with paragraph 5.d. of NASA Policy Directive, NPD-2220.4, "Use of the International System of Units (SI) in NASA Publications," a waiver has been granted for this report.

16. Abstract

Recent advances in understanding the physical phenomena controlling the interaction of electromagnetic energy with the ocean surface have revealed the possibility of remote measurement of the two-dimensional surface wave height spectrum of the ocean using bistatic radar techniques. The basic feasibility of such a technique operating at frequencies in the HF region (3 to 30 MHz) was examined during previous studies and hardware for an experimental verification experiment was specified.

The activities carried out to date have resulted in a determination of the required hardware and system parameters for both satellite and aircraft systems, the development, assembly, and testing of hardware for an experimental aircraft system, the development and initial testing of data processing procedures, and the conduct of an initial flight test experiment.

Activities during this phase were devoted to completing the assembly and testing of the experimental hardware, completing the experiment planning, conducting a field test experiment, and the processing and analysis of the experimental data.

During the experiments, the hardware operated satisfactorily; however, operational problems, low sea state during the test period, and noise on the 28 VDC aircraft power buss prevented acquisition of data suitable for generating directional spectrum maps of the test area or spectrum amplitudes.

Even though directional spectrum maps of the test area cannot be generated from the measured data, the hardware concept employed appears viable, and solutions to the problems encountered have been identified.

17. Key Words (Suggested by Author(s)):
Bistatic
Radar
Electromagnetic Energy
Hardware

18. Distribution Statement:
Unclassified - Unlimited
STAR Category 35

19. Security Classif. (of this report)	20. Security Classif. (of this page)	21. No. of Pages	22. Price*
Unclassified	Unclassified	58	

*For sale by the National Technical Information Service, Springfield, Virginia 22151

FOREWORD

This report covers activities performed by Battelle's Columbus Laboratories (BCL) on behalf of the National Aeronautics and Space Administration, Wallops Flight Center, under Contract No. NAS6-2006, "Services for Oceanography, Geodesy, and Related Areas Task Support". The NASA task monitor was Mr. H. R. Stanley. The Battelle program manager was Mr. A. George Mourad.

The investigation reported here was one of several tasks under the above-mentioned contract and represents the third phase of a program for developing a sea state monitoring system using a bistatic radar technique.

ABSTRACT

Recent advances in understanding the physical phenomena controlling the interaction of electromagnetic energy with the ocean surface have revealed the possibility of remote measurement of the two-dimensional surface wave height spectrum of the ocean using bistatic radar techniques. The basic feasibility of such a technique operating at frequencies in the HF region (3 to 30 MHz) was examined during previous studies and hardware for an experimental verification experiment was specified.

The activities carried out to date have resulted in a determination of the required hardware and system parameters for both satellite and aircraft systems, the development, assembly, and testing of hardware for an experimental aircraft system, the development and initial testing of data processing procedures, and the conduct of an initial flight test experiment.

The primary effort in this phase of the program was devoted to final assembly of the required experimental hardware, the designing and carrying out of an initial flight test, and subsequent data reduction.

The experimental system used consists of a surface-based FMCW transmitter radiating from 1 to 5 W. The transmitter sequentially radiates on ten frequency bands within the HF region. On each band, 256 sweeps over a 50-kHz range are performed during a 25.6 s interval. Each 50-kHz sweep is synthesized by stepping over this range using 5000 10 Hz steps. The receiving system is flown in an aircraft and synchronously follows the transmitter sweeps and band switches. Coherent Doppler processing of the received signal is made possible by the use of very stable basic frequency sources at both the transmitter and receiver locations. These sources are initially synchronized and maintain synchronization within the experimental requirements during the course of the experiment.

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by

G. T. Ruck, C. K. Krichbaum, and J. O. Everly

I. INTRODUCTION

This report covers activities performed by Battelle's Columbus Laboratories (BCL) on behalf of the National Aeronautics and Space Administration, Wallops Flight Center, under Contract No. NAS6-2006.

As one of the Tasks (Task II) under this contract, BCL had the responsibility for investigating the use of bistatic radar systems for the measurement of sea state. This investigation represents the third phase of a program for developing a sea state monitoring system utilizing a bistatic radar technique.

The bistatic radar approach involves the use of radio waves having a length parameter of the same order as the longer ocean waves contributing dominantly to the state of the sea, so as to produce a strong interaction. It has long been known that HF radio waves interact strongly with ocean waves. Significantly, the scatter is due to the Bragg effect, and its intensity depends directly on the heights of the ocean waves responsible for the scatter. By exploiting polarization, motion of the ocean waves, and motion of the receiver, a relative measure of the heights and direction of the dominant ocean waves can be obtained.

The basic bistatic configuration examined utilizes a surface-based HF transmitter located on a buoy or ship which is activated by command from an aerial platform or satellite. Radiation from the transmitter illuminates the nearby sea surface and is scattered toward the receiver. Both the direct and the sea-scattered signals are received. They are recorded or relayed to the ground for subsequent processing.

The basic feasibility of this technique was examined during the initial phase of this program, and the results reported in the final report on that phase issued in June 1972*. The results of the initial study indicated the potential of

*"Bistatic Radar Sea State Monitoring" NASA CR 137469, G. T. Ruck, D. E. Barrick, and T. Kaliszewski, Research Report, Battelle's Columbus Laboratories, Contract No. NAS6-2006 (June, 1972).

the technique and demonstrated the need for an experimental test of the concept. The activities during the second phase investigation were devoted largely to defining the equipment requirements for an aircraft experiment, developing, assembling, and testing the experimental equipment, and planning the experiment. The results of Phase II activities were reported in a final report submitted in May 1973.*

Activities during this phase were devoted to completing the assembly and testing of the experimental hardware, completing the experiment planning, conducting a field test experiment, and the processing and analysis of the experimental data. The activity during this phase and the results achieved are summarized in the following section. This is followed by a brief discussion of bistatic sea state measurement principles, a description of the experimental hardware, details of the conduct of the experiment, and a discussion of the data processing procedures. The report concludes with recommendations for further activities.

* "Bistatic Radar Sea State Monitoring", G. T. Ruck, C. K. Krichbaum, and J. O. Everly, Battelle's Columbus Laboratories, Contract No. NAS6-2006 (May 1973).

II. SUMMARY OF RESULTS AND RECOMMENDATIONS

Program Summary

The overall objectives of this study have been the determination of specific experimental hardware requirements, the acquisition, assembly, and testing of the required experimental hardware, and the design and execution of field test experiments utilizing an aerial platform for validation of the system concept. During this phase, final assembly of the experimental systems was accomplished. An initial field test experiment using an aircraft receiving platform was carried out, and the resulting data analyzed.

The specific objectives of the field test experiments planned for this phase were to determine the capabilities of the technique as a function of the system parameters, to validate the hardware approach, and to validate the data processing procedures, and algorithms. The flight test experiments were initially planned to be carried out in two steps. It was planned that measured data be taken during an initial test and this data processed to determine its quality. Any necessary modifications would be carried out and sufficient measurements then made to generate directional spectrum maps for the test area. As a result of delays in the acquisition of the required frequency synthesizers and scheduling problems with the simultaneous availability of the NASA aircraft, a helicopter for transportation to the Chesapeake light tower, and the NRL laser profilometer, it was possible to acquire only one set of measured data within the time and funding limits existing on this phase of the contract.

Results

The overall program to date has resulted in the following specific accomplishments:

(1) The major hardware requirements and system parameters have been identified for both satellite instruments and aircraft instruments for a bistatic configuration utilizing surface based transmitters.

(2) Experimental hardware for an aircraft system has been developed, assembled, and tested.

(3) An initial field test experiment has been conducted allowing the hardware concept to be evaluated and problem areas identified.

(4) Initial processing of the measured data has been carried out, and interface problems, with respect to A/D conversion of the analog data and subsequent digital processing, identified and resolved.

System and Hardware Specifications

Analysis of the experimental hardware requirements for an aircraft experiment reveal that Doppler processing to a resolution of 0.05 Hz is required. This is due to the much smaller geometric Doppler imposed on the received signal due to the aircraft's velocity than the Doppler that would result when a satellite is used. This requires that the transmitter and receiver be capable of maintaining a relative coherency of the order of 1 part in 10^9.

A system concept meeting all the requirements is a linearly swept FMCW waveform capable of range resolutions of the order of 3 or 6 km and Doppler resolutions of the order of 0.04 Hz. The required waveform can be synthesized using commercially available digitally controlled frequency synthesizers.

The digital sweep controller-programmer units to generate the required control signals for these synthesizers each utilize approximately 100 medium scale integration (MSI) TTL integrated circuits, and in effect constitute special purpose minicomputers. A broadband transmitter uses the frequency synthesizers as a driver. A transmitting antenna assembly with automatic switching of the required matching networks was assembled and mounted on the Chesapeake Bay light tower for the experimental test.

For the receiving system an automatic band-switching preselector and mixer assembly was utilized in conjunction with an R-390 HF communication receiver. The mixer local oscillator signal was supplied by one of the frequency synthesizers under control of the receiver digital sweep controller-programmer. The communications' receiver was tuned to a fixed IF frequency.

Directional Spectrum Measurements

During the experiments, the hardware operated satisfactorily; however, operational problems, low sea state during the test period, and noise on the 28V. D.C. aircraft power buss prevented acquisition of data suitable for generating directional spectrum maps of the test area or spectrum amplitudes.

Even though directional spectrum maps of the test area cannot be generated from the measured data, the hardware concept employed appears viable, and solutions to the problems encountered have been identified.

These problems can readily be solved by isolating the receiving system from the aircraft 28V. line, providing a portable data recorder independent of the recorder used for the laser profilometer, and repeating the measurements under higher sea state conditions.

Recommendations

It is recommended that an additional effort be undertaken as soon as practicable. The following specific steps should be carried out at that time.

(1) Modify the present experimental configuration as required in light of the information obtained during the preliminary aircraft flight test experiment.

(2) Carry out further flight tests during periods of high sea states.

(3) Reduce and analyze resulting data.

(4) Modify the system as required to permit operation by NASA personnel.

III. BISTATIC SEA STATE MEASUREMENTS

As discussed in the Phase I report, the average HF power scattered from an incremental area of the ocean surface for grazing illumination or observation is proportional to the power spectrum of the sea surface displacement. Thus, sampling at a number of wavelengths in the HF region is equivalent to sampling the magnitude of the surface energy spectrum at various wavenumbers. It should be noted that in contrast to most other wave measurement devices a bistatic instrument does not measure wave heights but the surface spectrum. Therein lies much of the strength as well as the weakness of the bistatic approach. In order to obtain the RMS surface height the integral of the spectrum over all wavenumbers must be taken. Obviously, the spectrum cannot be sampled at all wavenumbers and in order to obtain the RMS surface height with reasonable accuracies the portion of the spectrum not being sampled must contribute negligibly to the total potential energy of the surface. This implies that the sampling frequencies used should range from below the "knee" of the spectrum as illustrated in Figure 1 to well into the saturated portion of the spectrum. According to the Phillips' model*, the "knee" or peak of the spectrum should occur at a wavenumber uniquely related to the mean surface wind velocity. Due to the relationship between the radar wavenumbers and the surface spatial wavenumber for the bistatic geometry, this means that the peak of the spectrum for a given mean surface wind velocity will correspond to a particular illumination frequency. This is illustrated in Figure 2. From this we see for example that for a wind velocity of 10 knots the peak occurs at 17.6 MHz while for 5 knot winds the peak occurs at 70.7 MHz. Thus unless a very large frequency range is sampled the RMS heights corresponding to relatively calm seas, due to correspondingly low wind velocities, cannot be obtained.

On-the-other-hand, for 25 knot winds the theoretical cutoff occurs at 2.8 MHz and a system sampling between 3 and 30 MHz would be quite satisfactory since at 30 MHz the contribution to the surface energy is down by a factor of approximately 10^4 from that at 3 MHz, and frequencies above 30 MHz would contribute negligibly to the integration for the RMS surface height.

As an instrument strictly for the measurement of the RMS surface height, a bistatic system has the disadvantages indicated above. Its directional capabilities

* Dynamics of the Upper Ocean, Phillips, O. M., Cambridge University Press, London (1966).

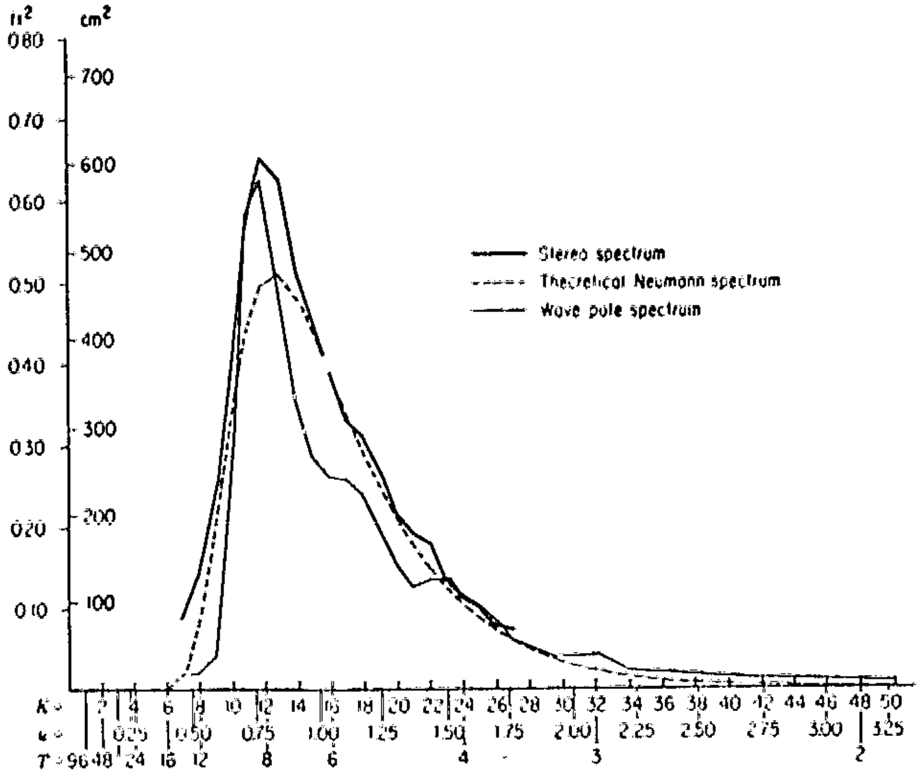


FIGURE 1. TYPICAL WIND WAVE MAGNITUDE SPECTRUM

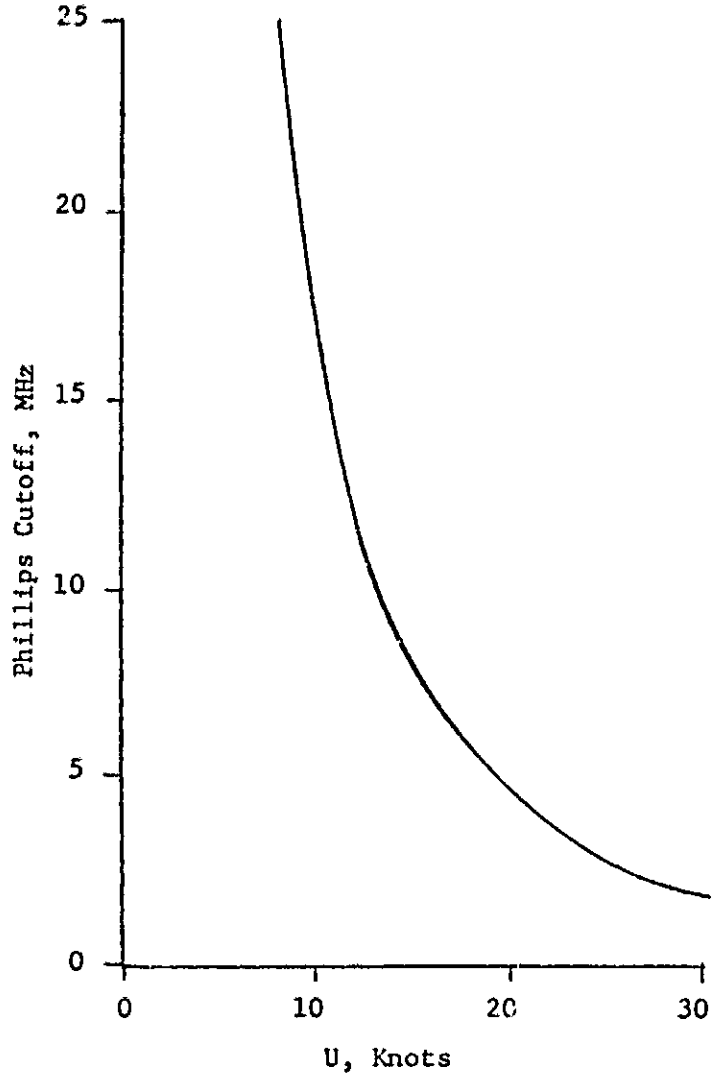


FIGURE 2. SPECTRAL PEAK IN EQUIVALENT RADAR FREQUENCY VERSUS WIND SPEED

of course are not required and its ability to measure very low sea states or very high sea states with a single instrument is limited.

It is, however, one of the few techniques available for directly measuring the directional spectrum rather than inferring it from surface height measurements and it provides the ability to select that portion of the energy spectrum where data is desired. For example, for fully developed seas the ocean wave height spectrum in the equilibrium range has a k^{-4} dependence on the magnitude of the wavenumber in the saturated region. This is predicted by dimensional considerations and verified by experimental data.* The shape of the spectrum peak and the shape for wavenumbers below the peak value are not readily predictable. A similar situation exists when the sea is not fully developed or for a so-called "confused sea" which is a mixture of growing sea, decaying sea, and extensive swell. Few theoretical models are available for these cases, and insufficient experimental data exists to establish empirical relationships. A bistatic measurement system could be particularly useful in such cases in measuring these portions of the spectrum of special interest.

Thus, the utility of a bistatic instrument depends upon the intended application. For simply monitoring RMS surface heights a system constrained to operating between 3-30 MHz could determine saturated wave heights corresponding to wind speeds between about 10-25 knots. The corresponding RMS wave heights would range from 1-10 ft. Such a system could be very simple if directional information were not desired.

The instrument could be used to examine a particular desired portion of the surface directional spectrum with considerable accuracy even though the RMS heights could not be obtained from such data. This allows the examination of the oceanographically interesting parts of the spectrum such as the peak and longer wavenumber regions preceeding the peak.

Since the primary objectives of the field tests carried out during this phase of the bistatic system development were to validate the hardware and data processing concepts it was felt that the system parameters should be choosen to measure the portion of the surface spectrum about which the most information is available. This is the saturated region for a fully developed wind sea with its corresponding k^{-4} dependence on wavenumbers. Because of the k_o^4 dependence of the surface scattering cross section on incident illumination wavelength, the received surface scattered

* Dynamics of the Upper Ocean, Phillips, O. M., Cambridge University Press, London (1966).

signal should be independent of frequency in the saturation region after the appropriate compensation of antenna, transmitter, and receiver gain variations with frequency are made. This is illustrated in Figures 3-10 of the Phase I report.

Under these conditions, a transmitter power output of 5 watts provides adequate signal-to-noise ratios for the predicted noise conditions in the test region. Contrary to the case for a satellite system as discussed in the Phase I report, the receiving antenna used during these tests was a vertically polarized stub mounted below the aircraft fuselage. As a result, the horizontal component of the scattered field is negligible and the scattering cross section is given by equation 13 of the Phase I report or

$$\sigma_{VV} = 2\pi k_o^4 \,(\sin\theta_s - \cos\varphi_s)^2 \,\{w^- \delta(-) + w^+ \delta(+)\} \quad , \tag{1}$$

with θ_s near $\pi/2$ for most of the flight paths. The corresponding part of the surface spectrum contributing to the scatter is that of wavenumber k,

$$k = k_o \sqrt{1 + \sin^2\theta_s - 2\sin\theta_s \cos\varphi_s} \quad . \tag{2}$$

Assuming aircraft velocities in the 200-250 mph range, the Doppler associated with various surface regions due to the aircraft motion is small so high Doppler resolution is necessary if good angular resolution is desired. The system parameters chosen, a 0.1 Hz basic repetition period and a 25.6 second integration time, provide an unambiguous Doppler range of ±5 Hz and a basic Doppler resolution of 0.04 Hz. For flight paths of up to 10 miles offset from the transmitting site and flight altitudes of around 5000', the frequency offset due to the direct path delay ranges up to about 50 Hz. This can be compensated for by an offset in the receiving system local oscillator ramp of this amount. However, with the SNR available and IF bandwidths of several kilocycles or greater it should not be necessary to compensate for this offset.

To aid in the resolution of the directional dependence it was planned that flights in parallel with the wind direction and perpendicular to the wind direction be flown over the test area. Since a vertically polarized receiving antenna was used the polar angle for the scattering geometry should be near 90°, requiring that the range to the aircraft be larger than its altitude. To accomplish this the aircraft was flown at 5000' along paths offset at CPA by 5 miles and 10 miles from the transmitter site.

IV EXPERIMENTAL SYSTEM DESCRIPTION

For a satellite-borne HF bistatic sea state sensor, the orbital motion of the satellite and the resultant Doppler shift imposed on the scattered signal can be used in conjunction with range resolution to determine the directional dependence of the ocean surface spectrum. Doppler spreads of as much as 40 to 50 Hz can result, and Doppler processing to provide a 1-Hz resolution would allow the surface spectrum directionality to be obtained with about a 10° resolution under these conditions.

For a typical aircraft-borne receiver, however, the Doppler spread would typically be of the order of a few hertz. Thus, to achieve the same surface resolution would require Doppler processing to a resolution of the order of 0.04 to 0.05 Hz. Doppler processing to this order presents several problems and complicates the experimental equipment requirements.

It is necessary to utilize a coherent waveform during the experiment and carry out Doppler processing. This allows the unambiguous directional spectrum to be obtained, and validates the concept of using the signal Doppler for obtaining the directional spectrum. In addition such factors as variations in the transmitter power output and the transmitting and receiving antenna gains with operating frequency can be removed by referencing the scattered signal level to the direct signal level.

The FMCW hardware configuration used in the experiment consists of two digitally controlled HF frequency synthesizers. One is used to drive an amplifier providing a power output of the order of 5 W and constitutes the transmitting equipment. The other is used as the local oscillator for the receiving equipment. An off-the-shelf HF communications receiver supplied by NASA was used with this configuration. The two synthesizers constitute the most critical and expensive items in this configuration, with the rest of the required components, such as the transmitter power amplifier, receiver, etc., available either as off-the-shelf equipment or assembled at BCL. The recording equipment was provided by NRL personnel. The digital sweep controllers necessary to control and program the synthesizer switching so as to provide the required frequency sweeps were designed, constructed, and tested at BCL.

The basic FMCW system is relatively simple, in that the transmitter radiates a coherent waveform which is swept over intervals of 50 kHz at a basic sweep rate of 500 kHz/sec thus providing 10 sweeps per second. The frequency versus time waveform is essentially a sawtooth in shape. At the receiver, the local oscillator is synchronously swept over the same frequency interval at the same rate, except that the sweep starting points are delayed in frequency by an amount corresponding to the range delay between the transmitter and receiver. The receiver local oscillator is also offset by an additional fixed amount corresponding to a desired IF frequency. This is in the HF band, and a conventional communications receiver is used as the IF amplifier. The scattered signal, after mixing with the receiver local oscillator then has a spectral spread which corresponds to different ranges. At the receiver output, the signal for each sweep is recorded on analog tape. The number of sweeps recorded is determined by the desired Doppler resolution. For example, if 256 sweeps are recorded at a rate of 10 sweeps per second, a Doppler resolution of 0.04 Hz results. Typically, a 50 kHz bandwidth is swept 10 times per second for 25.6 s. If 128 received signal samples per sweep are obtained in an A/D converter, then sufficient data exists to generate 64 range cells with a 6-km resolution per cell and 256 Doppler cells with a resolution of 0.04 Hz per cell. These are generated digitally by the application of a Fourier transform row by row and then column by column to the 128 x 256 matrix of data consisting of the 128 samples per sweep and the 256 sweeps. Actually, to save processing time, only the desired range cells are processed. Thus, only a few of the columns resulting from the first row-by-row transform are transformed.

In terms of hardware requirements, the stability and spectral purity required of the transmitter and receiver local oscillator for the FMCW case are the same as for a pulse-Doppler system. For the FMCW waveform, however, no peak power problems exist and the IF bandwidth required in the receiver is comparable to that normally encountered in HF communications receivers. For example, with the waveform parameters quoted above, an IF bandwidth of only 1.28 kHz is required.

The FMCW approach does require good time synchronization between the transmitter and receiver, however. For example, a 1-μs timing error in the sweep start time corresponds to a 0.3-km range error. The basic hardware requirements for a FMCW system are two digitally controlled frequency synthesizers with a high spectral purity and a stability of the order of 1×10^{-9} capable of operating in the frequency

range from 3 to 30 MHz, and sweep controllers for these synthesizers. A transmitter power amplifier, capable of supplying 5 W or more over the range from 3 to 30 MHz, is also required. The transmitting antenna system consists of a broadband vertical radiator and associated matching networks which are capable of operating satisfactorily over the range from 3 to 30 MHz. A block diagram of the transmitting equipment is shown in Figure 3.

The receiving equipment consists of one of the synthesizers and its associated sweep controller, an antenna preamplifier, an automatically band switchable RF preselector-mixer and a conventional communications receiver. The output of the receiver is coherently reduced to a baseband frequency and is recorded on an analog tape recorder for subsequent A/D conversion. A block diagram of the receiving equipment is given in Figure 4. The transmitter sweep controller is pictured in Figure 5.

Figure 6 shows the control signal interconnections among the various modules which comprise the transmitter sweep controller and between these modules and the external equipment. The sweep controller has the following internal functional circuit modules:

(1) decade boards,
(2) clock boards,
(3) control board, and
(4) band selector driver board.

The sweep controller has the following external operating control functions, switches, and indications:

(1) address switch,
(2) load switch,
(3) write switch,
(4) step drive interrupt switch,
(5) synthesizer drive interrupt switch,
(6) matching network align switch,
(7) sync switch,
(8) frequency band register,
(9) initial frequency register,
(10) local band indicating lights, and
(11) remote band indicating lights.

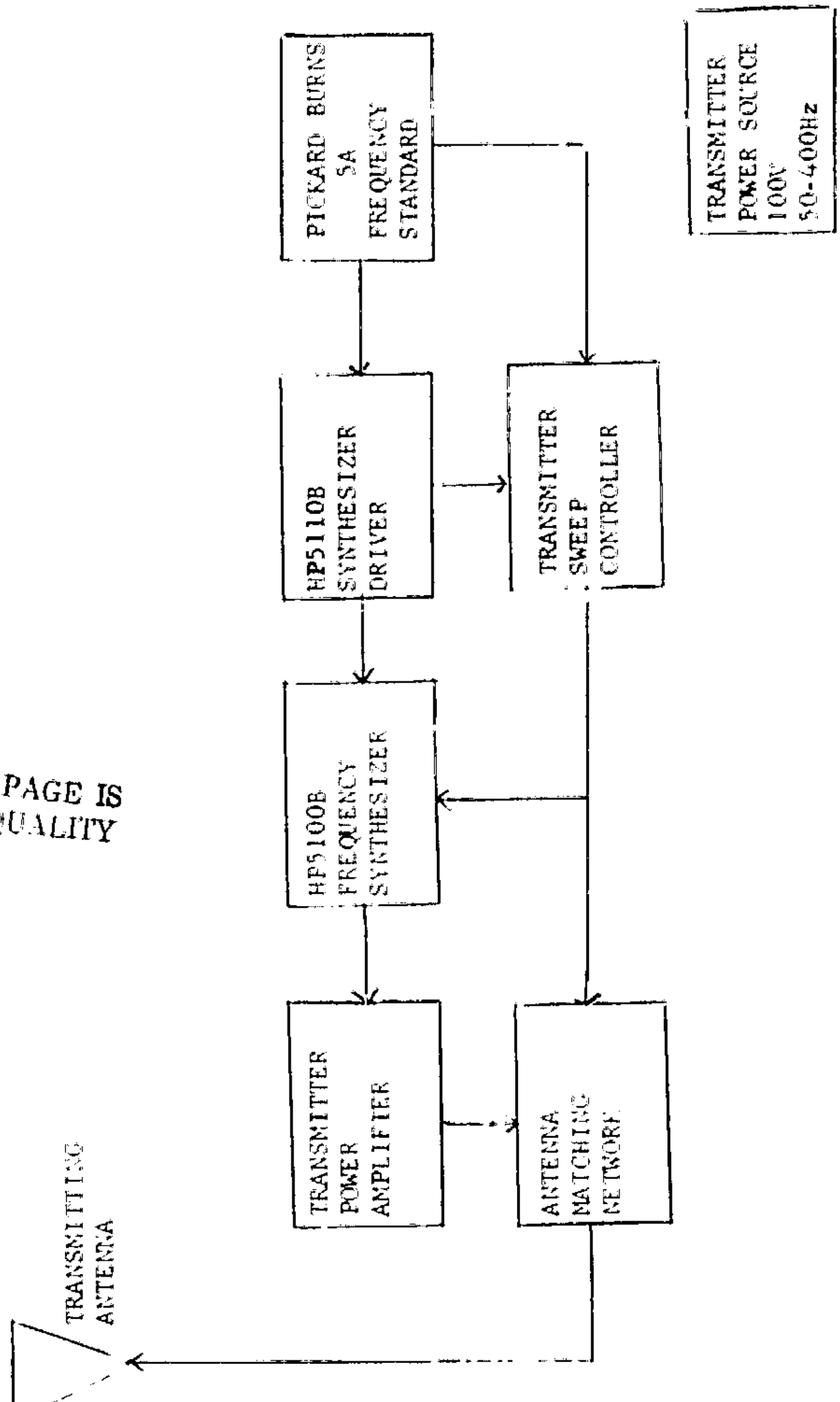


FIGURE 3. TRANSMITTING EQUIPMENT

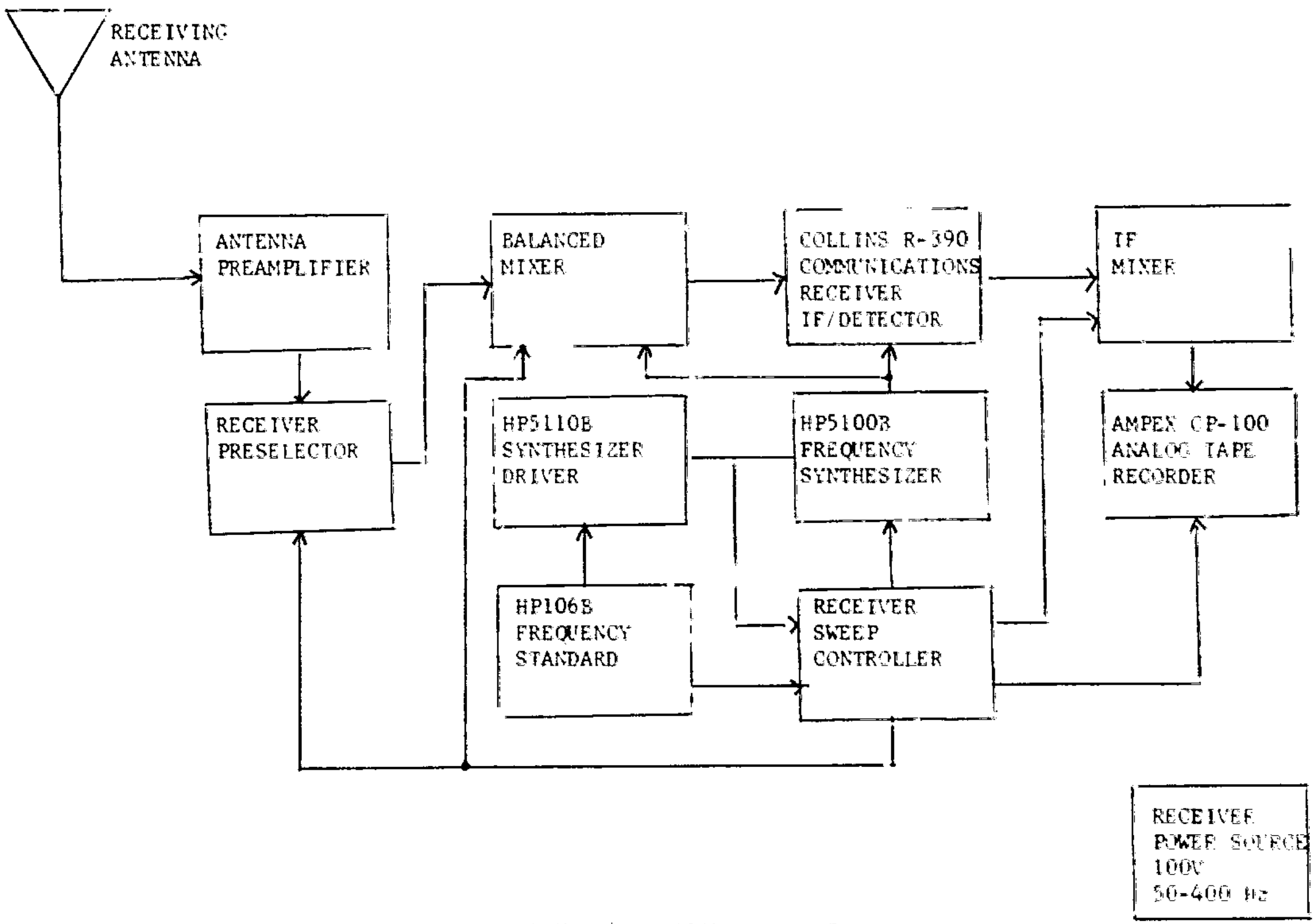
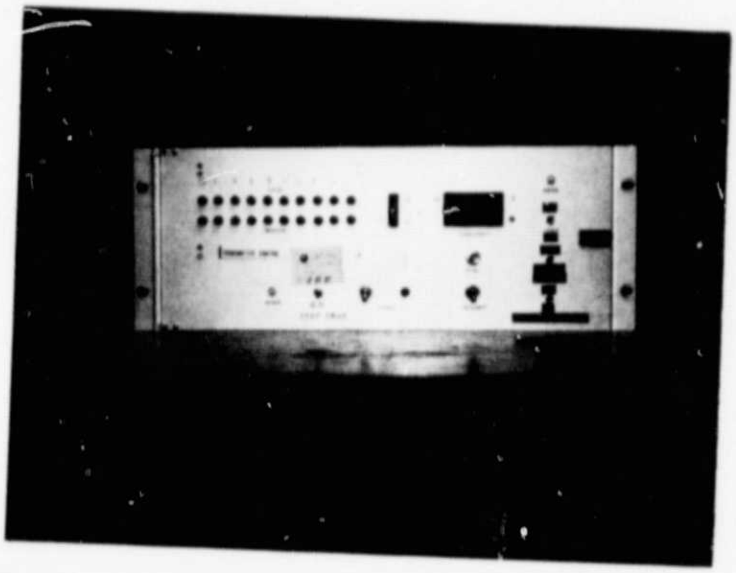


FIGURE 4. RECEIVER EQUIPMENT



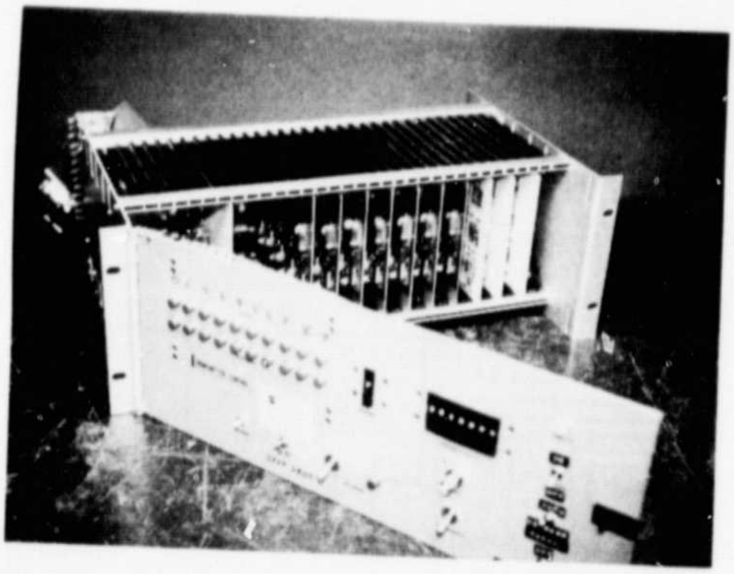


FIGURE 5. TRANSMITTER SWEEP CONTROLLER

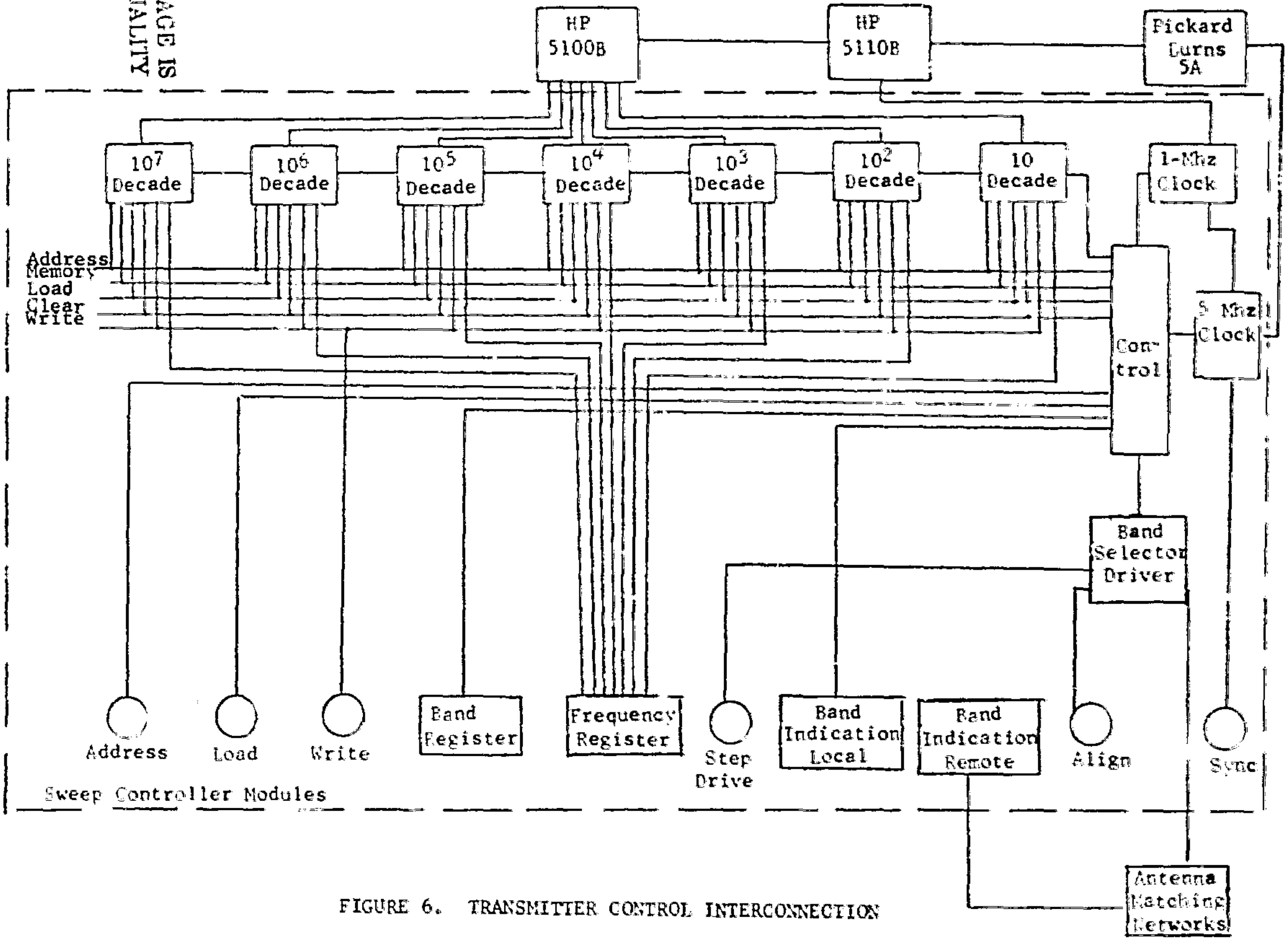


FIGURE 6. TRANSMITTER CONTROL INTERCONNECTION

The decade board is pictured in Figure 7. A schematic for the decade board circuit is contained in the Phase-II report.

Each decade board receives control signals from the control board, sweep-driving pulses from the adjacent decade board, and data input from the band and frequency registers. The ten output lines of a decade board drive directly the remote control inputs of one decade of the HP 5100 synthesizer.

There are two clock boards, a 1-MHz clock board and a 5-MHz clock board. The basic function of the 5-MHz board is to provide a master reset or synchronization pulse once each 320 seconds, i.e., after the sweep sequence has progressed over the ten different frequency bands. The 5-MHz board is driven directly by the 5-MHz sine-wave output of the frequency standard. At the transmitter the frequency standard is a Pickard Burns 5-A. At the receiver the frequency standard is an HP 106B. Since the two frequency standards are frequency-locked, the two 5-MHz clock boards are also frequency locked and the 320-second reset signals are in synchronism. The reset signals insure that the sweep at the transmitter is synchronized with the sweep at the receiver. The initial timing is achieved by physically removing the HP 106B from the receiver, taking it to the transmitter and connecting a sync line from the sync switch on the transmitter control panel to the receiver 5-MHz clock board which is permanently attached to the HP 1063 frequency standard. A similar sync line runs from the sync switch to the 5-MHz board in the transmitter sweep controller. The sync lines are attached to the "clear" terminals of the 5-MHz boards and the clear terminals are also connected through pull-up resistors to TTL ground. When the sync switch is activated both 5-MHz boards are cleared simultaneously, insuring that the reset signals will occur together as long as neither system is turned off. The 106B oscillator has a battery supply for use when the unit is moved. The 5-MHz clock board, installed inside the 106B at BCL, also has its own battery-powered power supply to keep it running during the course of the experiment.

A schematic for the 1-MHz clock board circuit is contained in the Phase II report. A schematic for the 5-MHz clock board circuit is contained in Figure 8.

The 1-MHz clock board is driven by a 1-MHz sine-wave output from the HP 5110B synthesizer driver. The latter equipment is driven by the 5-MHz sine-wave output of the frequency standard, so that the inputs to the 1-MHz clock board and 5-MHz clock board are phase coherent. The function of the 1-MHz board is to provide a wide range of digital frequencies to the control board for use in the sweep control sequence.

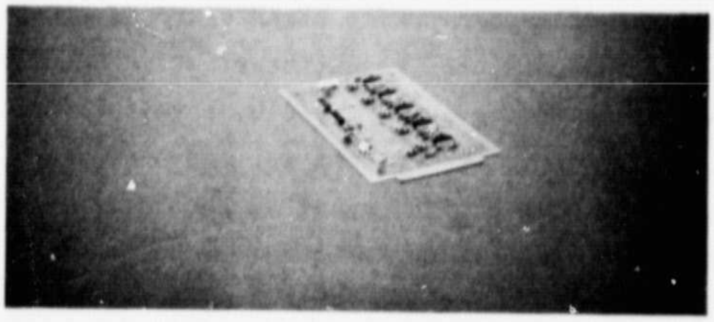


FIGURE 7. DECADE BOARD

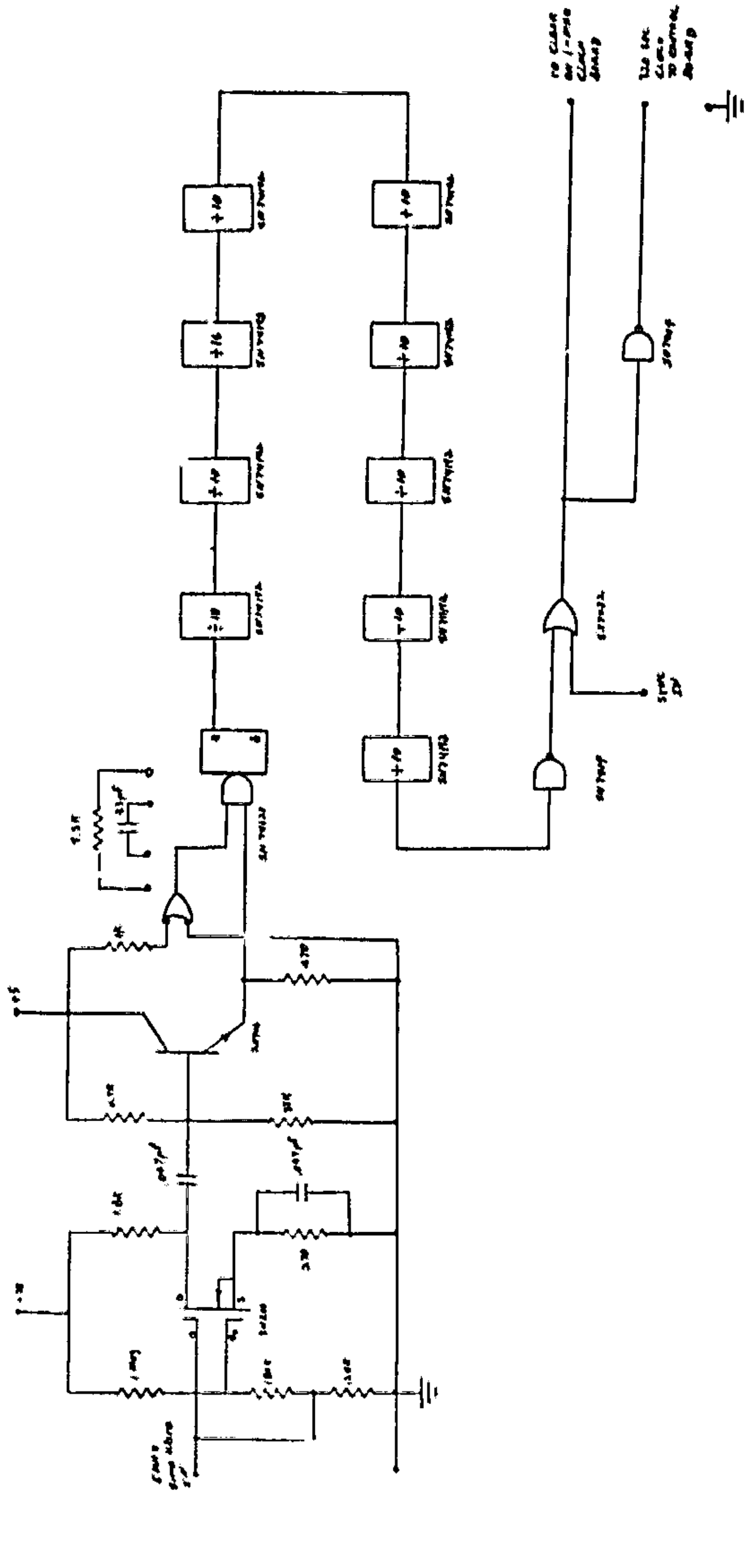


FIGURE 8. 5-MHz CLOCK BOARD

The control board is a sequential machine which uses the basic digital frequencies supplied by the 1-MHz clock board to generate a periodic sequence of voltage pulses and level shifts which control the operation of the decade boards and consequently the HP 5100 synthesizer and transmitted signal. In addition, control signals are supplied by the control board to the stepping motor which causes a rotary stepping switch to rotate and to switch in the proper matching network between transmitter power amplifier and transmitting antenna or the proper tuned circuits in the receiver preselector.

Figure 9 contains a logic schematic for the control board. The fundamental period of the control-board sequence is 30.4 sec, achieved by counting, with three cascaded shift registers (integrated circuits C1, C2, and C3), an input clock frequency of 1 pulse every 1.6 sec, generated on the 1 MHz clock board. The external flip-flop outputs of these shift registers make it easy to generate the transmitter enable signal consisting of an off-on-off-on sequence of 1.6 seconds, 1.6 seconds, 1.6 seconds, and 25.6 seconds. The transmitter operates on a constant frequency during the middle 1.6 second interval and is FM-modulated during the 25.6-second period.

The top waveform in Figure 10 shows the desired sweep as a function of time. The bottom waveform in Figure 10 shows the sweep achieved by using the HP synthesizer. The equivalent time (range) resolution of this waveform is approximately the reciprocal of the transmitted bandwidth, i.e., 20 μs. The Doppler resolution is approximately the reciprocal of the integration time, i.e., the duration of the sweep in a fixed band between F_1 and F_2. The sweep duration has been chosen to be 25.6 sec consisting of 256 sweeps each of duration of 1/10 sec. A number of 256 was chosen since it is the nearest power of 2 which will give the required Doppler resolution of 0.05 Hz. Having a number of samples which is a power of 2 is desirable since fast Fourier transform (FFT) algorithms used to process the measured data in range and Doppler, are most conveniently applied to data sets having dimensions of a power of 2.

The decade counter C4, serving both as a frequency register and address register, cycles through the ten states, 0 through 9, in response to the 30.4-second reset signal obtained from C3. The first of the two output functions of C4 is obtained by driving the local band indication lights on the programmer front panel by means of the BCD-to-decimal decoder C5. The second function is obtained by using the 4-bit BCD outputs directly to drive the 4 address lines which are connected to the seven decade boards. Each of the decade boards has a 64-bit random access TTL memory in which is stored the ten-digits corresponding to the ten start frequencies for the

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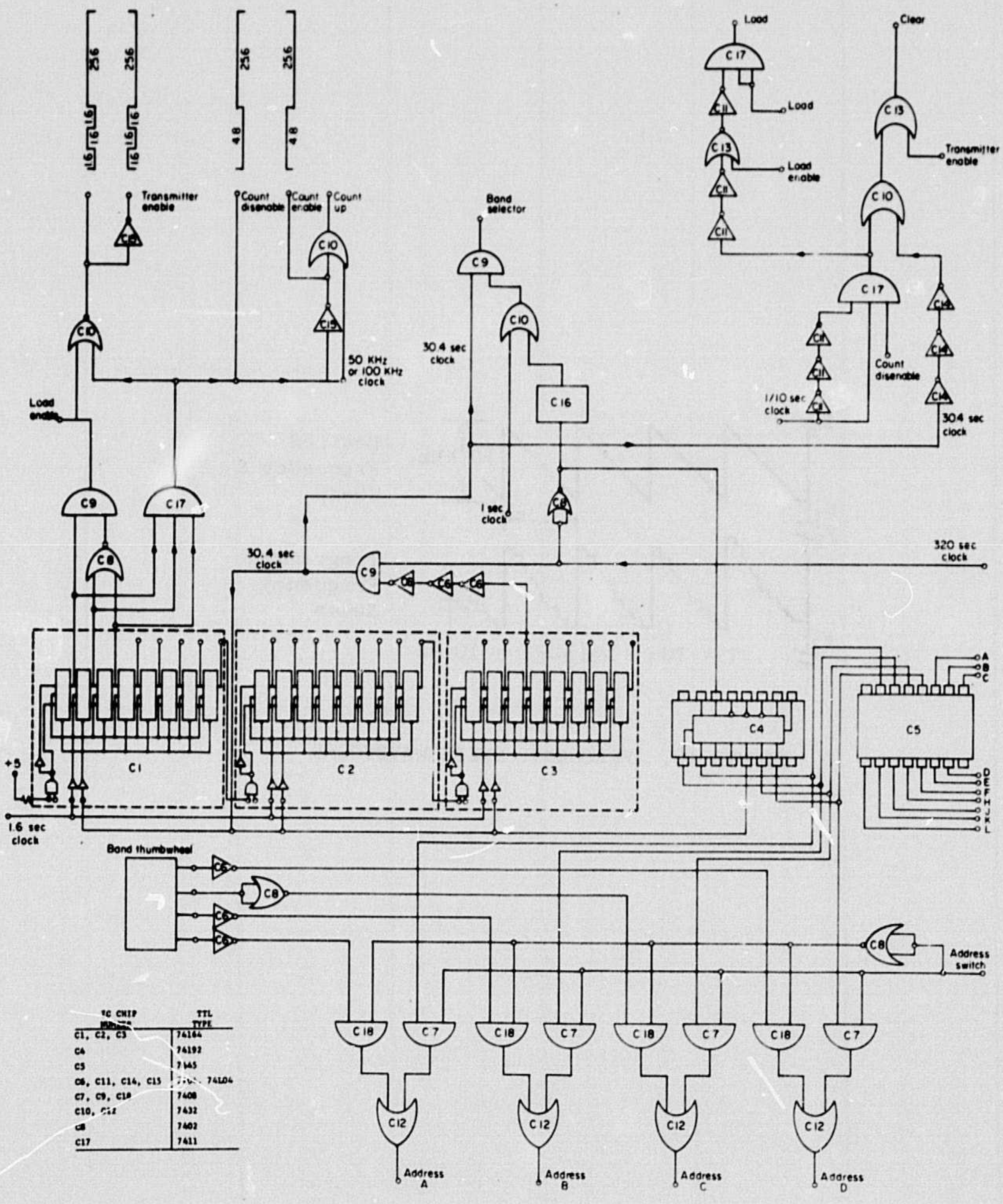


IC CHIP NUMBER	TTL TYPE
C1, C2, C3	74164
C4	74192
C5	7445
C6, C11, C14, C15	[illegible] 74L04
C7, C9, C18	7408
C10, C12	7432
C8	7402
C17	7411

FIGURE 9. CONTROL BOARD SCHEMATIC

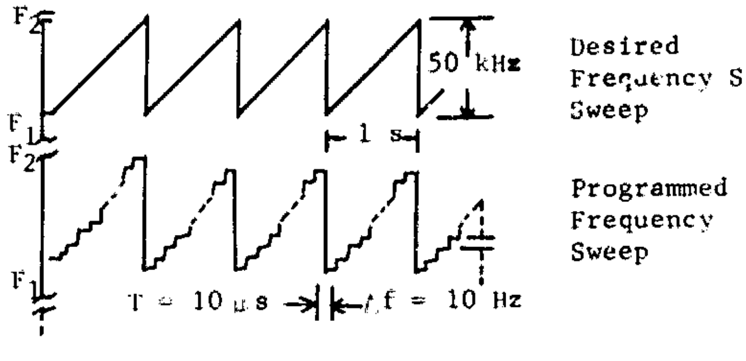


FIGURE 10. FREQUENCY SWEEP WAVEFORMS

ten frequency bands transmitted. The address lines set the address pointers in these memories to access the proper start frequency location according to which band is being swept. The combinatorial logic consisting of C7, C18, and C12 is used to switch control of the address lines between manual input from the band thumbwheel switch on the front panel and automatic input from C4. Manual input is used to load the start frequencies initially, and automatic input is used when the equipment is operating.

The circuitry at the upper-right-hand corner of Figure 9 is used to generate very short pulses on the clear and load buses to the decade boards. These clear and load pulses, slightly offset in time from each other, occur at the top of the frequency ramp, terminating the present sweep and loading the start frequency corresponding to the bottom of the frequency ramp to initiate the next sweep.

Circuit C9 provides a pulse to the relay driver board once each 30.4 s to move the stepping switch to the next position. The pulse-stretcher C16 and gate C10 supply two extra pulses each 320 seconds since there are two extra positions on the stepping switch which must be passed over after radiation on the tenth band is concluded.

Circuit C10 gates on and off the basic 50 kHz sweep frequency to the 10-decade board at the beginning and end of each 25.6 second sweep-on period.

The band selector driver board schematic is pictured in Figure 11. The input is an active-low pulse from the control board which is stretched in a 74121 multivibrator. The stretched pulse is used to turn on two transistor stages which apply approximately 24 volts to the coil of the stepping motor which turns the antenna matching network rotary stepping switch.

A brief description of the external operating control functions appearing on the sweep programmer panel follows. The address switch may be set for manual or automatic operation. Manual operation is required when loading start-frequency data into the decade board memories. Automatic operation is required when the system is operating in the sweep mode. The load switch is used to manually load a start frequency from the decade board memory into the decade board counter according to the address pointer set by the frequency band register. In this way the contents of each address can be examined. The write switch is used to write into the decade board memory at the address set by the frequency band register, the data digit set into the initial frequency register. The step drive interrupt switch suppresses the band selector drive pulses which rotate the rotary stepping switches automatically when the sweep moves from one frequency band to the next. This interrupt function is useful

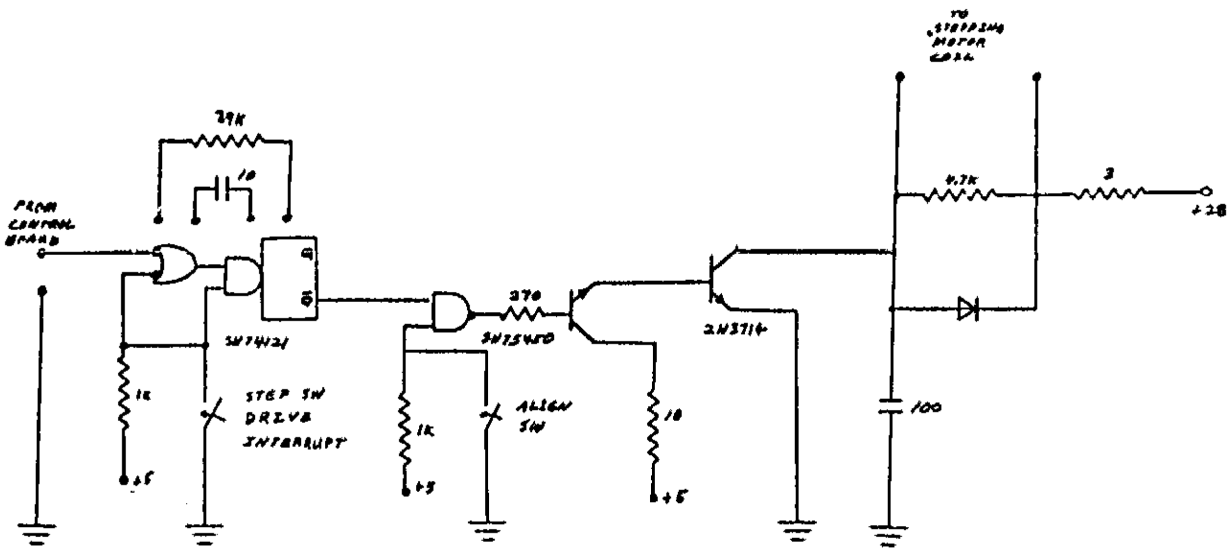


FIGURE 11. BAND SELECTOR DRIVER BOARD

when continuous radiation in a fixed frequency band is desired for calibration purposes. The matching network align switch can be used to manually position the rotary stepping switch to any desired frequency band. Each time this switch is depressed, the stepping switch rotates one position. The synthesizer drive interrupt switch may be used to suppress the drive voltage on the drive lines from the decade boards to the HP 5100B frequency synthesizers. The sync switch is used to align the two 5-MHz digital clocks as described previously. The frequency board register is a one-digit thumbwheel switch which can be used to access the decade board memory for manually writing or reading data into or from the memory. The initial frequency register is a seven-digit thumbwheel switch used for setting data into the decade board memories. There are two sets of frequency band indicating lights. The local band indicating lights give a readout of the address register on the control board. This tells in which band the transmitter or receiver electronics are operating. The remote band indicating lights indicate the current position of the antenna matching network rotary switch or preselector tuned circuit rotary switch. The two sets of lights give an instantaneous indication of whether the proper matching networks are in the transmitting or receiving signal path to the antenna for the current frequency band the transmitter or receiver local oscillator is operating in.

The receiver sweep controller is identical to the transmitter controller with the exception of a circuit board which contains three independent phase-locked loop frequency multipliers. Two of these are used to supply coherent signals to the R-390 receiver in-lieu of the local oscillators contained in the receiver, and the other drives a mixer which coherently beats the 455 kHz output of the R-390 IF amplifier to a 5 kHz coherent baseband data channel. These phase-locked-loops are driven by 500-kHz abd 5-kHz square wave signals generated on the 1 MHz clock board. Sinusoidal outputs of 11 MHz, 3 MHz, and 45 kHz are provided. A block diagram of the phase-locked-loop board is given in Figure 12.

The receiving antenna is a 36" vertical stub mounted in a port in the fuselage of the C-54 aircraft. A broadband antenna preamplifier is mounted at the antenna terminals to provide a high-impedance load for the antenna and a 50 Ω output impedance to match the coaxial cabling into the 50 Ω input of the preselector-mixer assembly. The antenna preamplifier uses an FET to provide a low noise figure and high dynamic range.

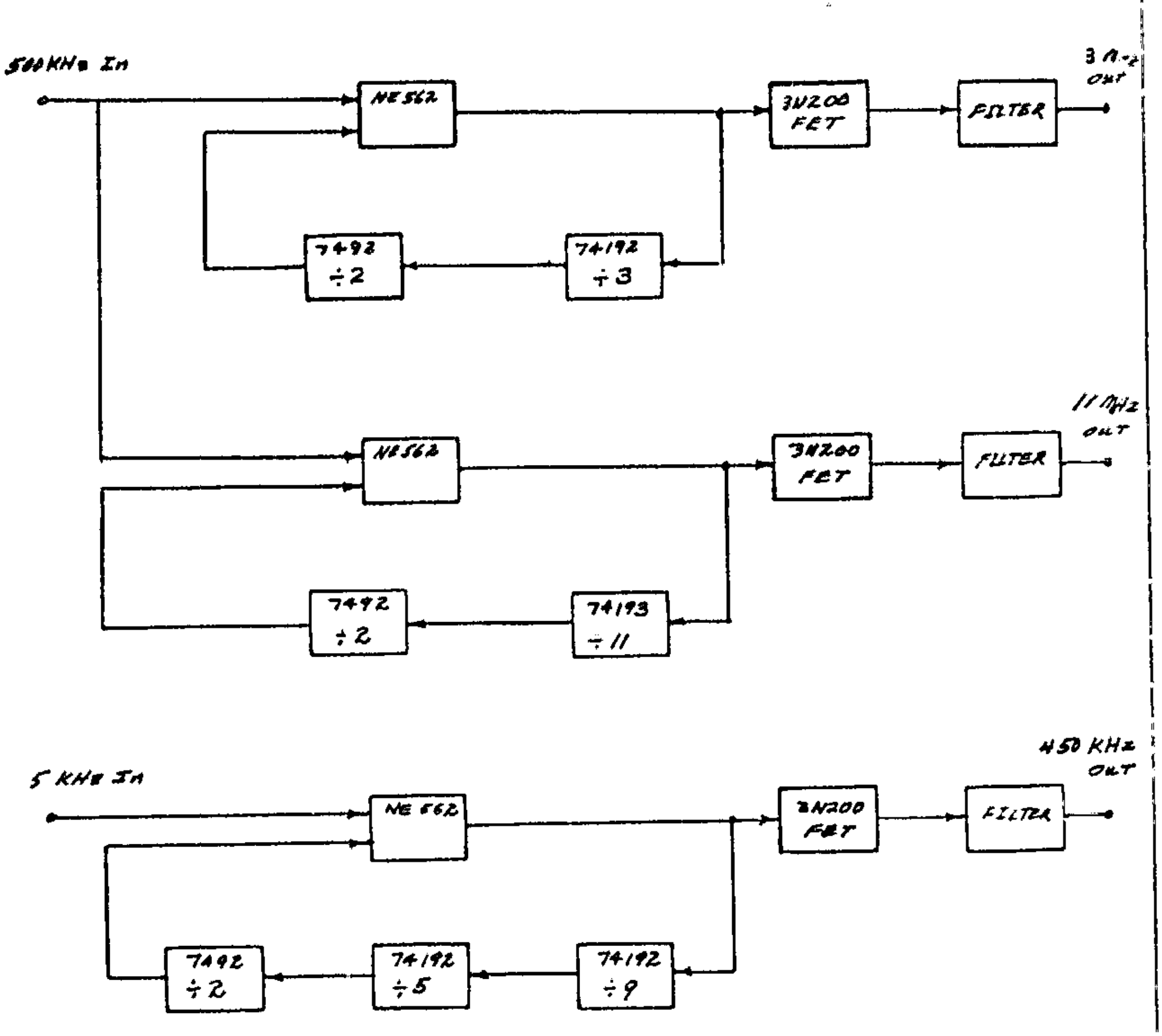


FIGURE 12. PHASE-LOCKED-LOOP BLOCK DIAGRAM

The 455 kHz IF output of the R-390 receiver is mixed with a coherent 450 kHz signal from the phase-locked-loop to provide a nominal 5 kHz coherent data channel. This data output is recorded along with a noncoherent output signal derived from the audio line of the R-390 with the receiver BFO on. This later signal is used primarily by the operator to insure that the transmitted signal is being received properly.

In addition to the above signal two reference signals are also recorded. One is the 5-kHz square wave from the 1-MHz clock board which drives the phase-locked-loop. This is used as a reference to allow recorder speed or frequency control variations to be compensated if required during the data processing. Another signal recorded is a timing signal from the 1-MHz clock board which is used to control the A/D conversion process and identify the start of the 0.1 second sweep periods.

System Initiation

The following procedure is required to start up either the transmitter or receiver system. Initially, the 5-MHz sine wave from the frequency standard and the 1-MHz sine wave from the HP 5110B driver must be connected to the sweep controller. The control board will begin the sweep sequence. The 1-MHz input signal must be removed during any 25.6-second interval of the sweep. This is at present necessary because the clear bus must be low during the reading of data into the system memory. The clear bus can be low only if the transmitter enable signal is low. This is low only during the 25.6-second interval. If, after disconnecting the 1-MHz signal from the controller input, a master reset signal from the 5-MHz clock occurs before the data read-in process is completed, the 1-MHz signal must then be reconnected for at least 4.8 seconds to insure the necessary clear low condition. To verify this condition, pin 12 of the control board can be checked for a TTL low condition.

The synthesizer drive switch should be off during the loading in of data and the address switch should be set to manual. Then, for each possible address digit 0 through 9 set into the frequency band register, a seven-digit data number can be written into the decade board memories by depressing the write switch once for each setting of the frequency band register. After all ten values of data are loaded, the contents of the memories can be checked by turning on the 5100B synthesizer, displaying its output on an oscilloscope, and turning the synthesizer drive switch on.

The specific frequency residing in a memory location can be set into the synthesizer and viewed on the oscilloscope by setting the frequency band register to the frequency band number and moving the load switch to on. When all ten frequency bands have been verified, the system is placed in the sweep mode by turning the load switch off, moving the address switch to automatic and reconnecting the 1-MHz input. The align switch then can be used with the aid of the local-remote indicating lights to move the matching network rotary switch to the proper position corresponding to the band the sweep programmer is operating in.

V. EXPERIMENT CONDUCT

As described in the previous section, the experimental system consists of a surface based FMCW transmitter which sequentially transmits on ten frequencies in the HF band, and an airborne receiver system whose local oscillator synchronously sweeps with the transmitter. A necessary initial step in the use of this system is to synchronize the sweep starting times of the transmitter and receiver units. This is done by means of a portable clock assembly for the receiver constructed from an HP-106B portable frequency standard and a TTL digital master clock driven from this standard. A similar nonportable transmitter master clock is driven by a Pickard-Burns frequency standard of relatively unknown specifications. The two units are synchronized by physically and electrically bringing them together and then not allowing power to be removed from either unit after that. The portable clock is returned to the receiving location after synchronization.

After primary power has been applied to both the transmitting and receiving equipment the sweep starting frequencies must be loaded into the memories of the control units. This is done by a front panel thumb wheel switch which selects the specific channel and a seven decade switch which allows the frequency to be selected to the nearest 100 Hz. When power is removed from either of these units the frequency data must be reloaded into the memory.

The transmitting frequencies used during the test were:

Band	Frequency
0	3.15 MHz
1	4.15 MHz
2	5.15 MHz
3	6.7 MHz
4	7.15 MHz
5	8.53 MHz
6	12.15 MHz
7	15.15 MHz
8	19.15 MHz
9	25.15 MHz

The R390 receiver which functioned as an IF amplifier for the receiving system was set to 8 MHz, so the sweep start frequencies for the receiver synthesizer were set 8 MHz above those for the transmitter.

An initial attempt to obtain experimental data was carried out during the week of March 25-29, 1974. The transmitting equipment was installed on the Coast Guard operated Chesapeake Bay light tower located at 75° 42.8' West longitude and 36° 54.02' North latitude. The tower is pictured in Figure 13. The tower deck where the transmitting antenna was installed is located 90' above mean sea level and is 60' square. The average water depth in the tower area is 50'. Installation and check out of the transmitting equipment on the tower was completed on March 26. The NASA helicopter which transported the transmitting equipment and BCL personnel to the CPB tower crashed while returning to the tower to transport the personnel back to Wallops Station and they were forced to stay overnight and return to Wallops the following day (March 27) via Coast Guard cutter and subsequent land transportation. Installation of the receiving equipment in a NASA Wallops C-54 aircraft was initiated on Thursday, March 28 and continued on Friday, March 29. The receiving equipment installation and check out was not completed due to the unavailability of 28V D.C. primary aircraft power at the equipment racks of the C-54. As a result, no data flights were made due to lack of 28V power and the unavailability of another helicopter qualified for overwater flying to transport the portable clock to the CPB tower for synchronizing the transmitting and receiving units. In addition the NRL laser profilometer for ground truth verification was not available for the next several weeks.

BCL personnel returned to NASA, Wallops Station during the week of April 25-26, the earliest time at which both the aircraft and laser profilometer were available after acquisition of another helicopter by NASA. The receiving installation in the C-54 was completed and ground checked on April 23. On April 24, the portable clock was transported to the CPB tower by helicopter and synchronized with the transmitting clock. A calibration flight was made during the afternoon of April 25. No useful data was obtained at this time due to a malfunction in the receiver-mixer assembly and a malfunction in the NRL instrumentation recorder used for data recording.

On Friday, April 26, a series of flight paths were flown and data recorded. In-so-far as could be observed during the flight, the bistatic equipment appeared to be functioning properly. The flight paths flown consisted of three calibration circles around the transmitting site at an altitude of approximately 500' at a radius

FIGURE 13. CHESAPEAKE BAY LIGHT TOWER

of ≈ 1 mile. During these flights the transmitter was not sweeping and was transmitting pure CW at 3.15 MHz, 8.53 MHz, and 25.15 MHz in succession. Following this, three along-wind flights were made over the tower, 5 miles offset from the tower, and 10 miles offset from the tower at an altitude of approximately 5000'. Each path was flown for approximately 10 minute period. A similar series of paths was flown in the cross-wind direction. The system was sweeping properly during these flights.

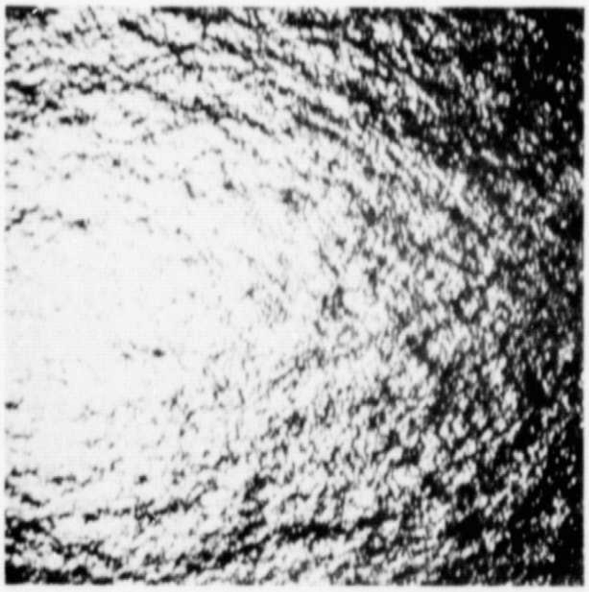
During this period, the surface winds were from the SW at an average speed of 8 knots. The wave heights in the area of the tower were visually estimated to be 1-2' with 3' of swell running north. Figures 14a and 14b illustrate surface conditions. These were taken from 5000' and 500' respectively. In figure 14a the aircraft shadow can be seen in the upper left hand corner.

As a result of the low waveheights no useful laser profilometer data was obtained, thus the measured data cannot be validated.

During the data flights one difficult was encountered. The recorder used, an NRL Ampex CP-100, was part of the laser profilometer system. Four channels of data output as described in the previous section were recorded using this recorder. Due to the unavailability of the direct record electronics for this recorder, FM recording was used. The recording speed required to provide a 5 kHz bandwidth for the bistatic coherent data output is 15 ips. Drive belts were not available for speeds in excess of 3-3/4-ips, thus NRL personnel set the recorder for 3-3/4 ips which provided a maximum bandwidth of 1.25 kHz. To compensate for this and bring the coherent data into this range the receiver local oscillator offset was adjusted to place the data frequencies in excess of 3.75 kHz away from the receiver IF center frequency and thus within the passband of the recording system. There was no way however of retaining the 5 kHz reference frequency information, and the stability of the record and playback system had to be wide enough to accommodate the offset thus increasing the noise and interference somewhat.

Among the operational difficulties encountered during the experiment was the requirement for removing the portable receiving clock assembly from the aircraft after each flight and transporting it to a location where it could be plugged into 115V. A.C. primary power to prevent interruption of operation. This unit is relatively heavy and requires two men to remove it from and install it in the aircraft. At one point during this process the unit was dropped by the NASA technicians installing it.

(a) From 5000' altitude



(b) From 500' altitude

FIGURE 14. OCEAN SURFACE AT TEST SITE

The unit was operating at the time and continued to operate with no apparent damage. Unfortunately, this occurred after the unit had been synchronized with the transmitting clock and any effects on the synchronization could not be ascertained.

Another problem which manifested itself during ground checks on the receiving equipment after installation in the aircraft was severe noise on the 28V. D.C. primary power line of the aircraft. In general, among other things, this made it very difficult to adjust the lock-in range for the receiver phase-locked-loops and tended to cause spurious locks. At the time this was attributed by NASA personnel to the ground power unit which was powering the aircraft and assurances were given that the aircraft 28V. power buss was not subject to this noise during flight. This was not the case as will be subsequently discussed.

Another problem encountered during flight was the loading of the sweep starting frequencies into the receiver controller. Loading of seven digit numbers for all ten frequencies has to be accomplished within a five-minute interval between master clock clears and while flying no simple means is available for verification that the correct frequencies are loaded into the memory. It turned out to be distressingly easy to get an incorrect frequency loaded into the memory and a verification mechanism is needed.

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VI. DATA REDUCTION ALGORITHMS

The data processing required to obtain a directional spectrum from a set of measured data consists of those steps necessary to obtain a range-Doppler map for the scattering area, and those steps necessary to convert the range-Doppler information into directional spectrum information.

Range-Doppler Processing

Figure 15a shows the variation with time of the transmitted and received signal frequencies for the FMCW bistatic radar. The basic time delay R/c is due to propagation delay, where R is the total range and c is the speed of light. If the receiver local oscillator frequency is as shown in Figure 15b. The resulting signal to be processed has an effective pulse repetition period of T_r, where T_r is the period of the transmitted frequency ramp, and an effective pulse width of T_r-R/c. The target information is contained in a band centered about frequency f_1. Thus, frequency shift is equivalent to time delay and in order to effect range (or time) resolution; a frequency decomposition of the received difference-frequency signal must first be obtained. This is most easily carried out digitally, so that a discrete representation of the recorded receiver voltage must first be obtained by means of analog-to-digital (A/D) conversion.

A frequency decomposition of the set of discrete samples obtained from an individual pulse repetition period provides an effective pulse shape, i.e., a distribution of received scattered signal energy versus time (or equivalent range). This pulse shape or range interval can be subdivided into range cells to give the system range resolution, which is proportional to the reciprocal of the transmitted bandwidth. No Doppler information regarding the motion of the receiver platform or that of any intervening reflective surfaces can be obtained by examining one pulse period. The Doppler information is contained in the variation with time of the scattered signal energy within the same range cell, i.e., by a frequency decomposition of the data vector made up of the signal energy of the same range cell observed over a consecutive sequence of pulse periods. The integration time is the time duration of the sequence of pulse periods. The reciprocal of this integration time is the Doppler resolution of the system.

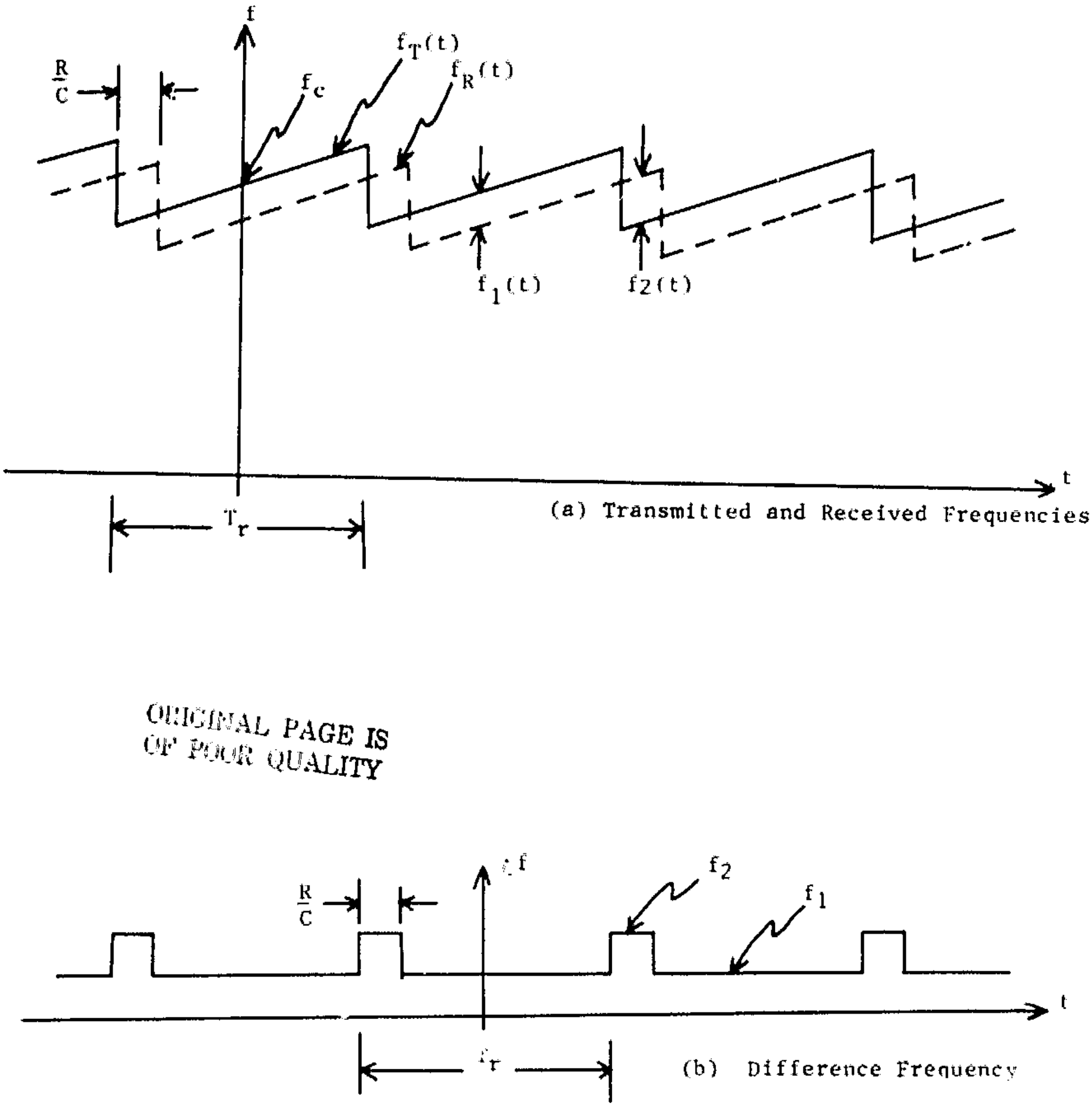


FIGURE 15. FREQUENCY OF TRANSMITTED AND RECEIVED SIGNALS

Figure 16 shows the normalized A/D conversion of the receiver voltage recording for approximately 4 seconds of time during the recent aircraft experiment. This represents the data for about 40 consecutive sweeps of the 256 recorded at 3.15 MHz. The recorded analog data was run through a bandpass filter prior to A/D conversion to minimize aliasing problems. The passband of this filter was about 5 to 125 Hz. The sample rate was 500 samples per second. Figure 17 shows the auto-covariance function computed from the data of Figure 16 (notice that considerable noise exists in the signal). Figure 18 shows the Fourier transform of the autocovariance, and thus represents a spectrum or frequency decomposition of the receiver voltage recording of Figure 16. Consequently, Figure 18 represents an average pulse shape or range interval averaged over 40 pulse intervals. The slow variation of signal energy from pulse interval to pulse interval has been averaged out and no Doppler information is present. In order to obtain Doppler information, precise timing information with respect to the starting point and stopping point of each individual sweep or repetition period of the data of Figure 15 is required. A distinct frequency decomposition would be required for the data samples contained within each repetition period. If then a matrix of data were obtained whose row vectors are the frequency decompositions or spectra of a set of consecutive repetition periods, the column vectors of this matrix represent time samples of signal energy variations of distinct range cells. A further frequency decomposition of column vectors for range cells of interest would represent the Doppler spectra for these range cells.

Directional Spectrum Processing

To obtain the spectrum information from the range-Doppler maps requires that a number of geometric transformations be carried out. Initially the direct signal level must be identified from the range cell containing it and used to normalize the signal levels in the other range cells. For path geometry such that the direct path is greatly different from the scattering path in arrival angle at the receiver the effects of receiving and transmitting antenna vertical radiation patterns must be included. For the antennas used this amounts to a $\cos^2$ dependence. In general this will be necessary only for paths near the transmitter zenith. For the

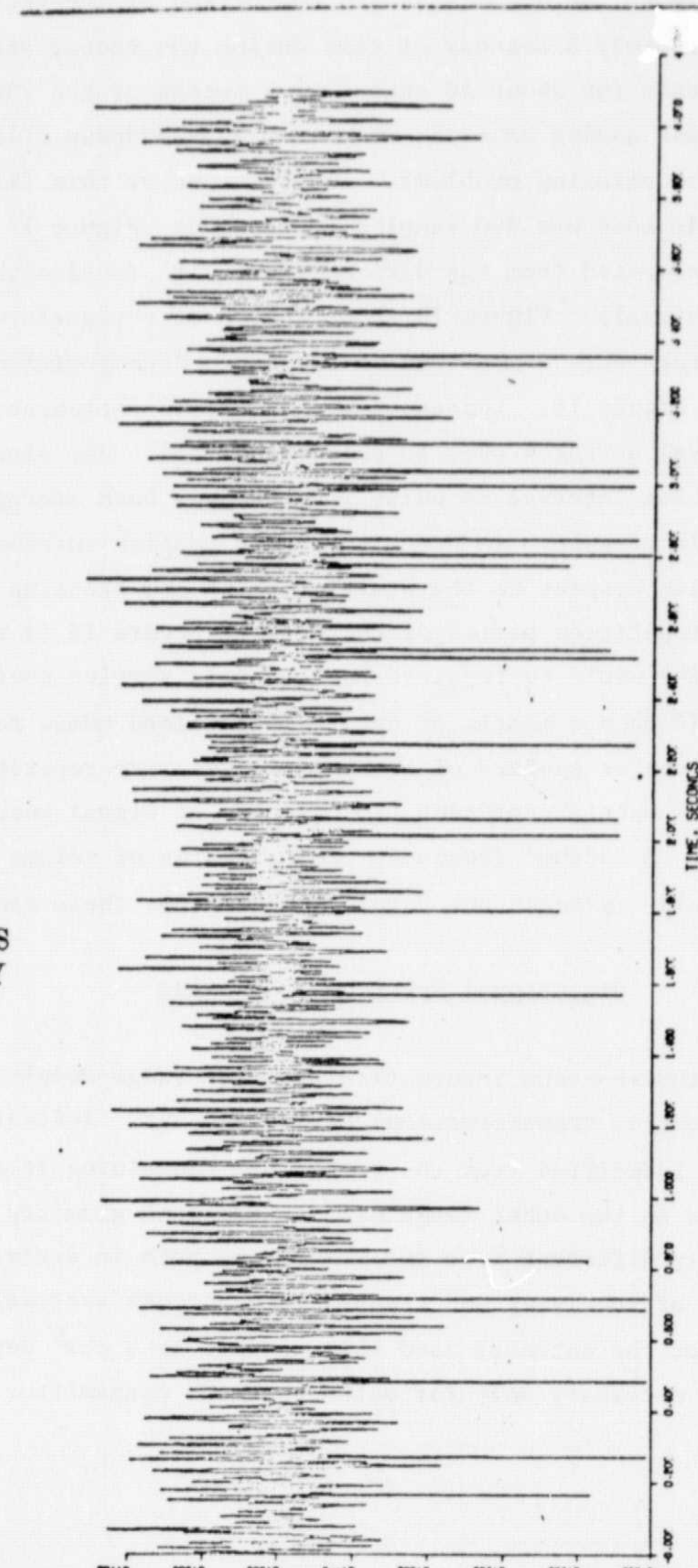


FIGURE 16. RECEIVER VOLTAGE RECORD

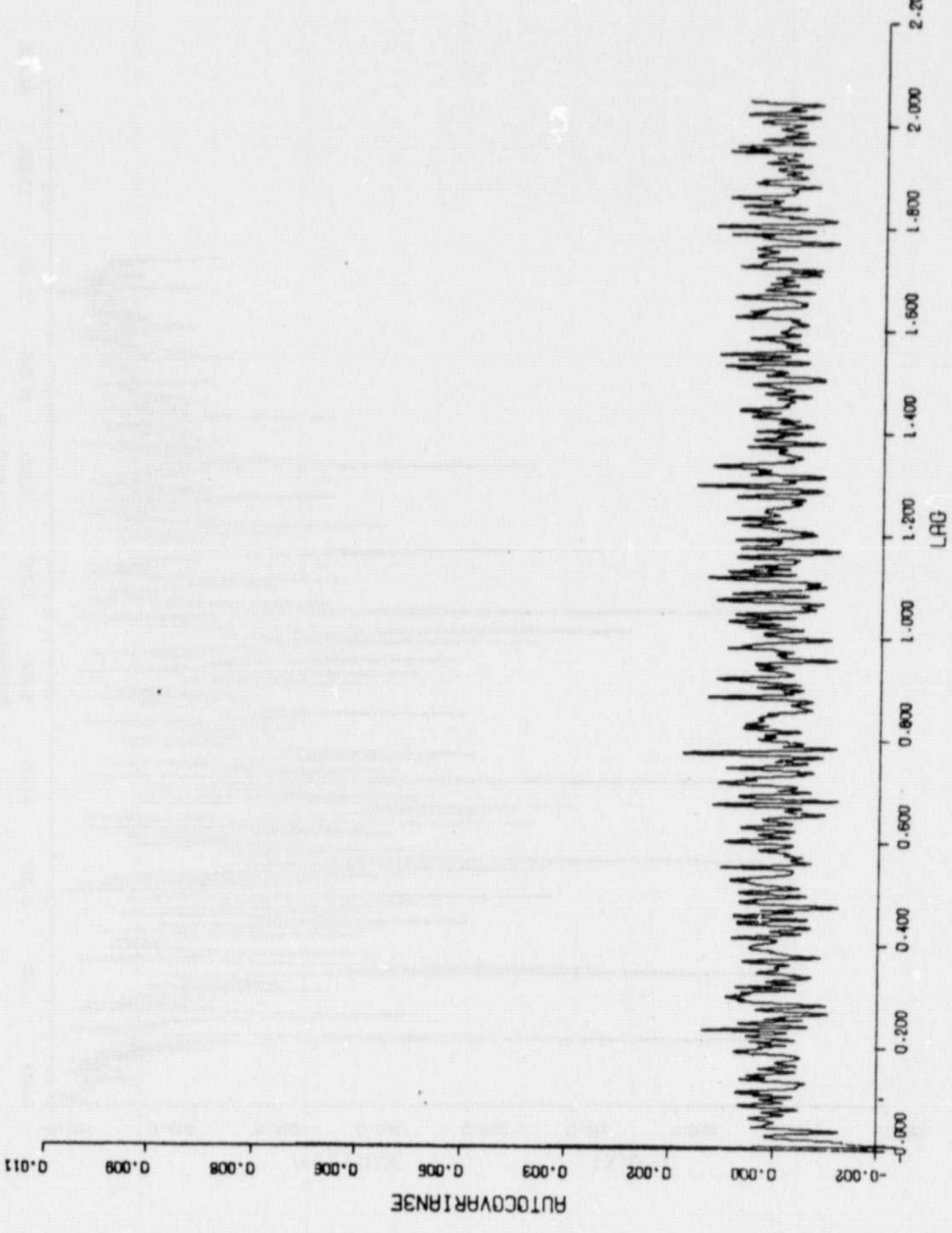


FIGURE 17. AUTOCOVARIANCE FUNCTION OF THE DATA OF FIGURE 16

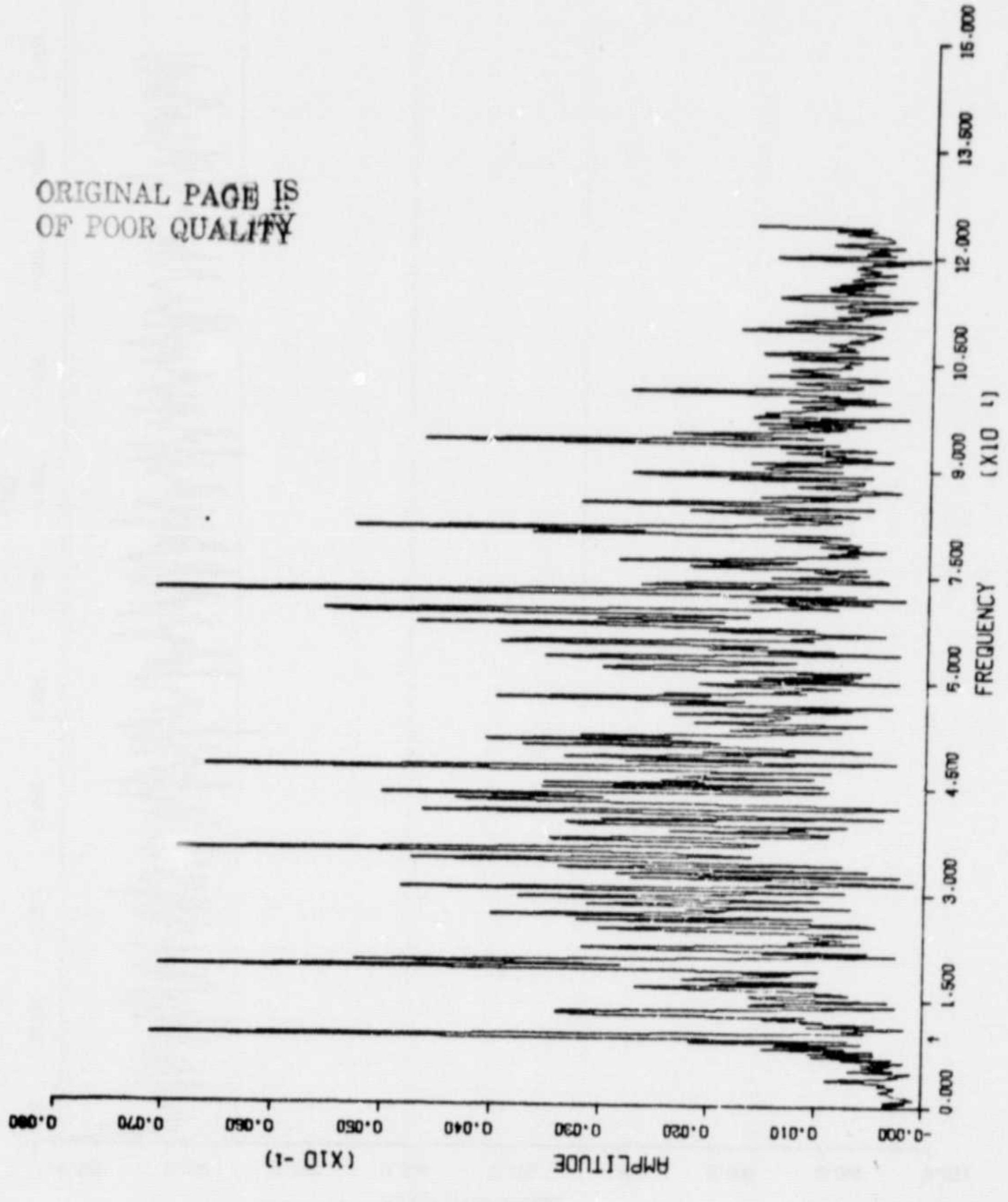


FIGURE 18. AMPLITUDE SPECTRUM OF THE RECEIVER VOLTAGE RECORD

data taken at 5 mile and 10 mile offsets these angular effects can be neglected without significant error and the direct path signal level used directly to normalize the data for other ranges. This must be done for each sample frequency.

For a particular range cell, corresponding to a known delay time τ, the Doppler frequency can be used to locate the scattering region on the surface relative to the aircraft. This is required to determine the geometric scattering angles so that equation 1 for the scattering cross section can be inverted and the spectrum value obtained. Unless tracking data is available for the aircraft, the aircraft's velocity, the time relative to CPA (Closest Point of Approach), the aircraft range at CPA, and its altitude must be known. The range at CPA can be obtained directly from the measured data, as can the time relative to CPA. The aircraft's velocity and altitude must be known during the data flights.

Given the geometry as illustrated in Figure 19, the range to a scattering region on the surface as a function of the delay, the polar angle for the aircraft, and the azimuthal angle for the scattering region is given by

$$\rho = \frac{2hc\tau + c^2\tau^2 \cos\theta'}{2c\tau\cos\theta' + 2h\left[1-\sin\theta'\cos(\varphi-\varphi')\right]} \quad . \tag{3}$$

The azimuthal and polar angles θ', φ' of the aircraft relative to the transmitter are given by

$$\theta' = \cos^{-1}\frac{h}{R_0} \tag{4}$$

and

$$\varphi' = \pi/2 \sin^{-1}\left\{\frac{VT}{R\sin\theta'}\right\} \quad , \tag{5}$$

where R is the range, h the aircraft altitude, V the velocity, and T the time from CPA. T is negative before CPA and positive after.

A table of surface ranges to the scattering area, ρ, is computed for values of φ ranging from 0 to 360°. These values of φ and the corresponding values of ρ are used to compute the differential Doppler from each scattering region.

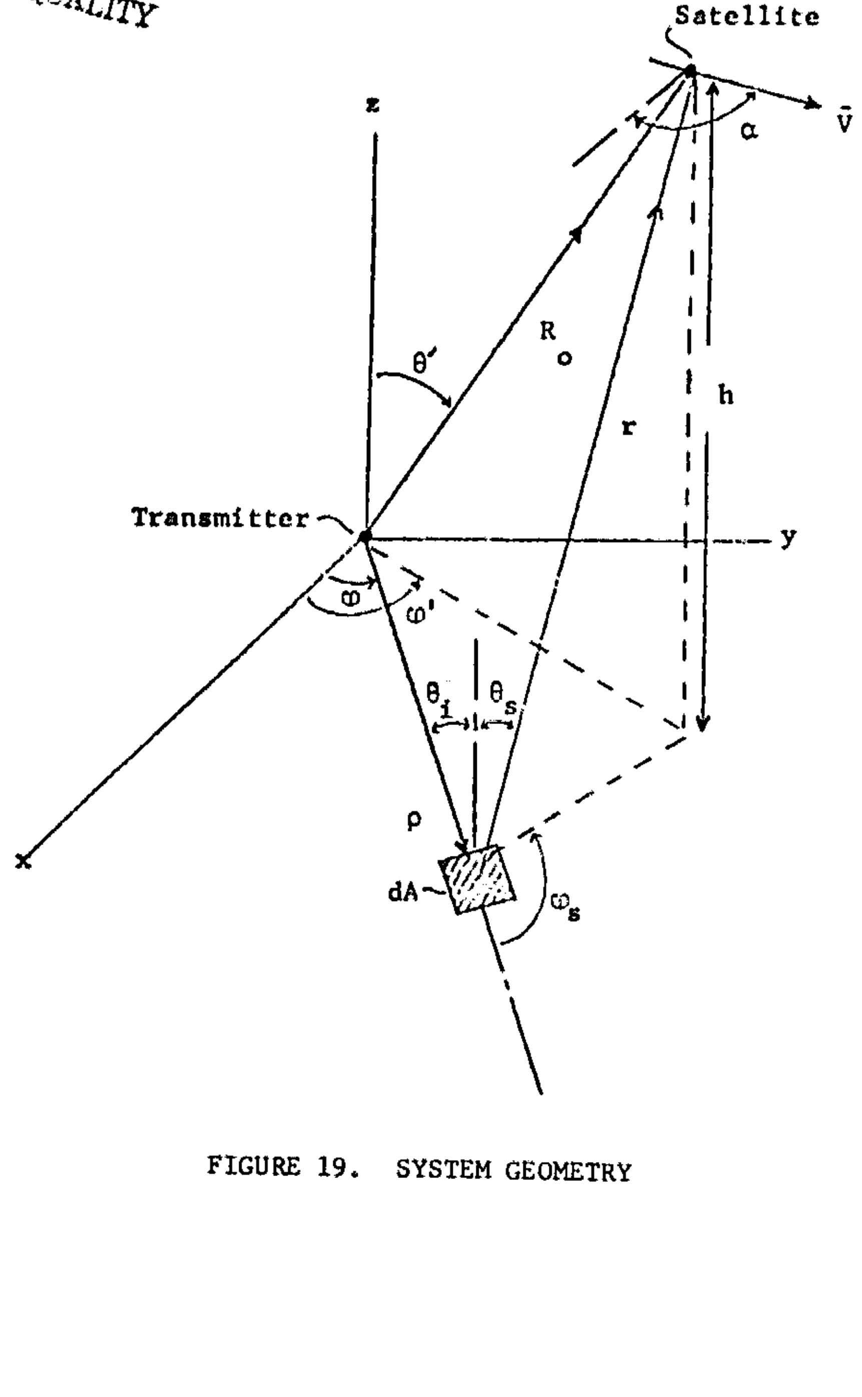


FIGURE 19. SYSTEM GEOMETRY

$$f_d = V\frac{f_o}{c} \qquad \frac{\rho\ \cos\varphi - h\sin\theta'\ \cos\varphi'}{\sqrt{h^2\ \rho^2\cos^2\varphi' - 2\rho h\ \cos\varphi'\sin\varphi'\ \cos(\varphi'-\varphi)}} \tag{6}$$

Due to the ambiguity associated with the $\cos(\varphi'-\varphi)$ term a specific Doppler value is associated with two different values of φ. These are separated giving for a particular delay τ tables of f_d versus φ. These are inverted to obtain the two values of φ associated with each differential Doppler frequency f_d. This assigns a spatial position ρ, φ to each Doppler component for a specific range cell.

The scattering parameters for each spatial cell are given by

$$\varphi_s = \sin^{-1}\left\{\frac{R\sin\theta'\ \sin\ (\varphi' - \varphi)}{\sqrt{\rho + R^2\sin^2\theta' - 2\rho R\sin\theta'\cos(\varphi-\varphi')}}\right\}, \tag{7}$$

and

$$\varphi^s = \cos^{-1}\left\{\frac{R\cos\theta'}{\sqrt{R^2+\rho^2-2\rho R\sin\theta\ \cos(\varphi'-\varphi)}}\right\} . \tag{8}$$

These are calculated for the center ρ,φ of each spatial surface region and used in conjunction with equations 1 and 2 to solve for the spectrum magnitude and the associated vector wavenumber $\overline{k}$. To obtain the correct scattering cross section magnitude associated with each spatial region the change in area associated with a given resolution cell $\Delta\tau\Delta f_d$ must be taken into account. This requires that the Jacobian representing the transformations from τ, f_d coordinates to ρ,φ coordinates be calculated as a function of ρ and φ. The value of $|J|$ required is given by

$$\frac{\partial\rho}{\partial\tau}\frac{\partial\varphi}{\partial f_d} - \frac{\partial\rho}{\partial f_d}\frac{\partial\varphi}{\partial\tau} . \tag{9}$$

The partial derivatives required for this are given by

$$\frac{\partial\rho}{\partial f_s} = \frac{\lambda_o}{V}\left[\frac{\cos\theta'\cos\varphi}{D} - \frac{\rho\cos\theta'\cos\varphi - h\sin\theta'\cos\varphi'}{D^3}\right.$$

$$\left.(\rho\cos^2\varphi' - h\cos\theta'\sin\theta'\cos(\varphi-\varphi'))\right]^{-1} \quad , \tag{10}$$

$$\frac{\partial\rho}{\partial\tau} = c\left[1 + \frac{2\rho\cos\theta' - 2h\sin\theta'\cos(\varphi-\varphi')}{D}\right]^{-1} \quad , \tag{11}$$

$$\frac{\partial\varphi}{\partial f_d} = \frac{\lambda_o}{V}\left[\frac{\rho\cos\theta'\cos\varphi - h\sin\theta'\cos\varphi'}{2D^3}(2\rho h\cos\theta'\sin\theta'\sin(\varphi-\varphi')) - \right.$$

$$\left.\left(\frac{\rho\cos\theta'\sin\varphi}{D}\right)\right]^{-1} \quad , \tag{12}$$

and

$$\frac{\partial\varphi}{\partial\tau} = \frac{cD}{2\rho h\sin\theta'\sin(\varphi-\varphi')} \quad , \tag{13}$$

where

$$D = \sqrt{h^2 + \rho^2\cos^2\theta' - 2\rho h\cos\theta'\sin\theta'\cos(\varphi-\varphi')} \quad , \tag{14}$$

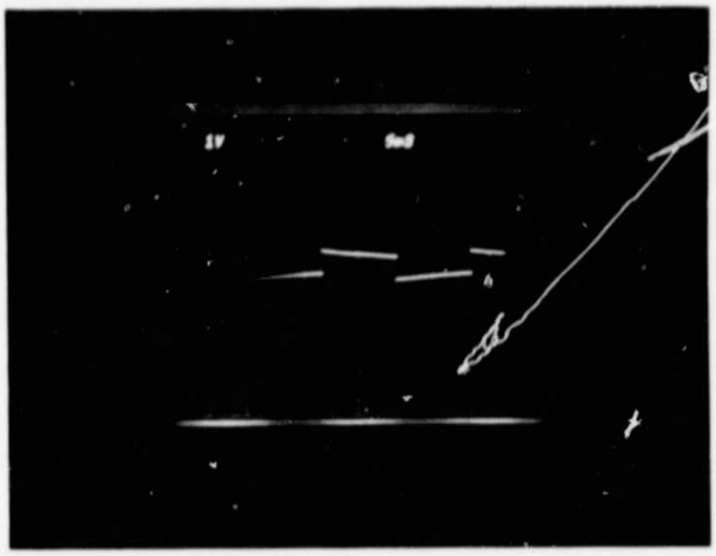
The signal value for each range-Doppler resolution cell is further normalized by the Jacobian calculated from the above equations to obtain the equivalent cross section for each cell. From this the spectrum value can be obtained. Each range-Doppler cell for each operating frequency yields a spectrum value corresponding to a particular spatial wave number magnitude and direction. This constitutes the directional spectrum estimate. Each range cell provides an independent set of directional spectrum estimates. In addition each five minute operating period provides additional independent data. Since the basic surface spectrum should not change significantly over the observation time or the range examined, these values can be averaged to improve the accuracy of the spectral estimates.

Experimental Data Processing

The data tapes collected during the experiment were returned to BCL and A/D converted and processed to generate the range-Doppler tables as indicated. It readily became apparent that some serious problems existed with the data. As indicated previously, sweep timing information in the form of a flip-flop output voltage waveform obtained from the 1-MHz clock board of the receiver system on the aircraft was recorded during the aircraft experiment. However, this timing channel was so contaminated by noise that it could not be used in the A/D processing. The magnitude of the noise problem can be appreciated by an examination of Figure 20. Here are shown two oscilloscope photographs of the sweep timing signal. The top photograph shows the trace obtained in the laboratory. The lower photograph shows a playback of the sweep timing signal recorded during the aircraft experiment. Two problems are immediately obvious: (1) there was a tape speed anomaly or incompatibility in one or both of the two tape recorders used to record and playback the data and (2) the recorded sweep control data channel has a noise component of sufficient amplitude so that no voltage threshold can be set such that this signal on playback can reliably trigger the Schmidt trigger control channel of the A/D converter.

The tape speed problem may be due to the fact that two separate instrumentation recorders were used for the record and playback functions. The NRL AMPEX CP-100 machine was used to record the data on the aircraft. An AMPEX FR-1300 was used at BCL to playback the data. It appears that the data was either not recorded at the speed indicated or the proper electronics for that speed were not used in the recorder.

The noise problem is due to the use of the aircraft 28V. D.C. power buss for the receiver system. The 28V. D.C. voltage is used on the 1-MHz clock board as part of the squaring circuit used to convert the input sine wave to a TTL-compatible square wave. This squaring circuit, also used on the 5-MHz clockboard, is shown in Figure 8. Any A.C. components present in the D.C. voltage are superimposed on the TTL signals and propagated throughout the clock board integrated circuits. The 28V. supply voltage is also used directly on the phase locked loop circuits in the receiver IF section and any power supply noise signals appear in the received voltage data. It is also used to power the receiver IF mixer and as a result this noise is induced directly on the data signals in the coherent output channel.



(a) Laboratory measurement of sweep controller

(b) Playback of data recorded during aircraft experiment

FIGURE 20. SWEEP TIMING DATA CHANNEL VOLTAGE OSCILLOSCOPE TRACES

Since the complete loss of sweep timing and data coherency prevent Doppler processing it was felt that perhaps the data could be range gated and a nondirectional amplitude spectrum generated. In general however, the noise in the data channel makes even this very difficult. It is possible that if the 10 Hz component and its harmonics due to lack of sweep timing information are removed, and extensive data smoothing used to eliminate the noise, amplitude data could be recovered. The results obtained after processing three frequency channels indicate however that the quality of the data obtained is not worth the time and expense involved. This is due not only to the spurious noise introduced by the noise on the 28V. power buss but by the degredation in SNR due to the slight surface roughness as well.

Even if the spectrum were to remain at its peak value for wavenumbers below the cutoff, the k_o^4 dependence of the scattering cross section causes a serious loss in signal-to-noise ratio for low sea states. For example, the cutoff point for an 8-knot wind is 27 MHz and at 3 MHz the k_o^{-4} scattering behavior would result in a 38 dB reduction in scattered signal level. It is obvious from Figure 1 that the spectrum decreases relatively rapidly for wavenumbers below the cutoff value so the SNR reduction at the lower frequencies is even lower.

In a confused sea such as existed during the measurement period, energy will generally be present in spectral regions below the wind sea cutoff. This is illustrated in Figure 21 which compares NRL significant wave height measurements over a number of years with predictions from the Phillips model. At the lower wind velocities the significant heights are consistently higher than predicted. This is largely due to the presence of decaying sea and swell propagated from other areas. Unfortunately, no analytical or experimental guidance is available to predict the behavior in these regions and in view of the lack of laser profilometer verification data and the noise present in the data, additional processing to recover these spectral amplitudes would be of little value.

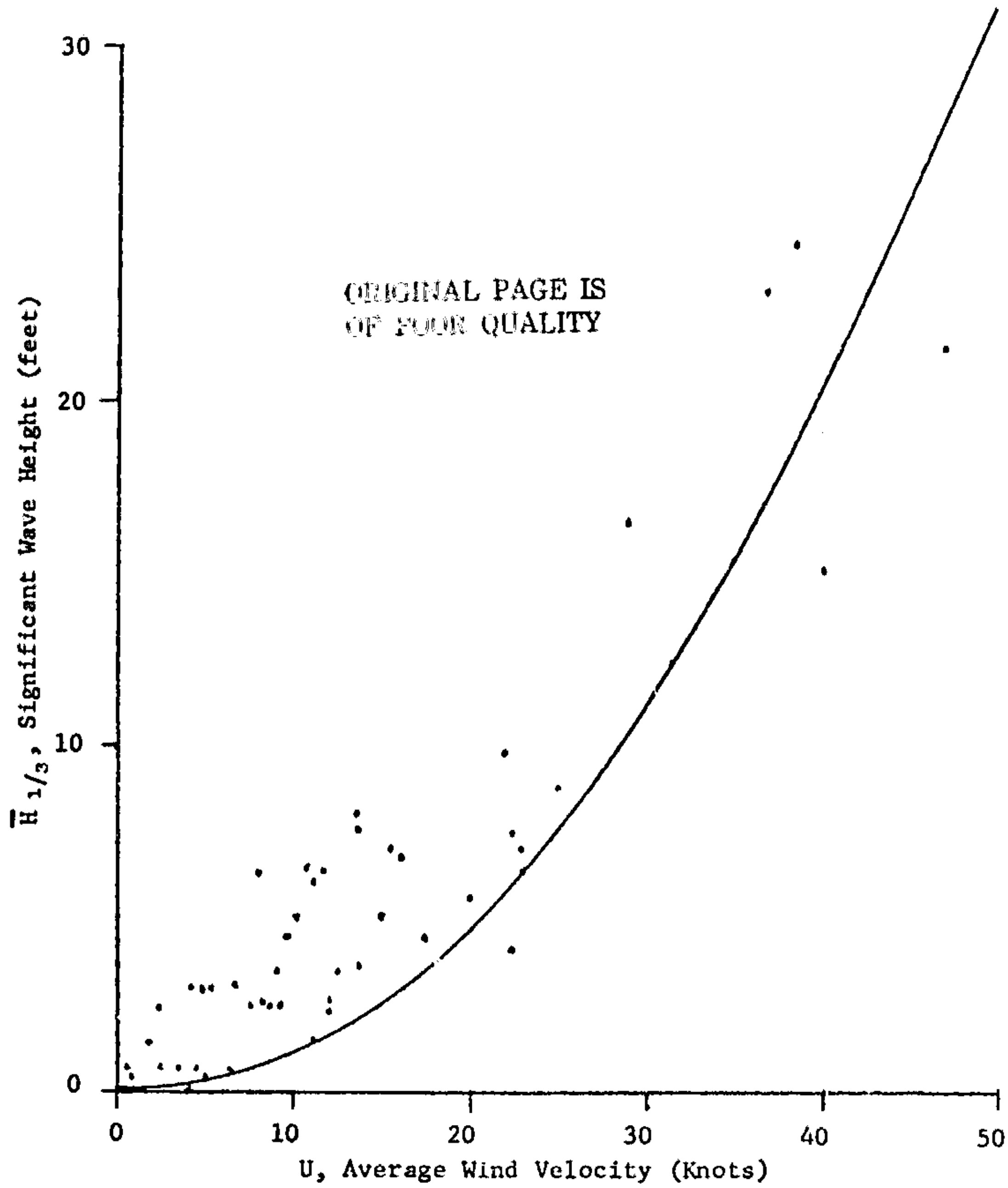


FIGURE 21. EXPERIMENTAL SIGNIFICANT WAVE HEIGHTS VERSUS THEORETICAL PREDICTIONS

VII. CONCLUSIONS AND RECOMMENDATIONS

Satellite System Specifications

In-so-far-as could be determined the hardware and data processing concepts utilized are valid and a satellite system with simpler hardware requirements should be feasible. The basic specifications for such a satellite system are given in Table 1.

TABLE 1. SATELLITE SYSTEM SPECIFICATIONS

Transmitter

Power Output - 10 watts average
Frequency - 3 to 30 MHz, switched in 1 to 2 MHz steps
Stability - 1 part in 10^8
Spectral width - < 0.5 Hz
Pulse width - 10 to 20 μs
Antenna - vertical polarization, azimuthally omnidirectional
PRF - 500 to 1000 pps

Receiver

Frequency - 30 to 30 MHz, switched in 1 MHz steps
Noise figure - not critical
Local oscillator stability - 1 part in 10^8
Time synchronization - not critical
IF bandwidth - 50 to 100 kHz
System bandwidth - [illegible] to 50 Hz
Coherent Integration Time - 1 s
Range Gates - 3 or more
Antenna - Horizontally polarized

It appears that a pulse-Doppler system would be satisfactory. A direct pulse system would require a duty cycle of 50 to 200, and pulse compression may be necessary to reduce the peak power requirements. This could be either a phase coded system or FM chirp. The horizontal receiving antenna at the satellite could consist of an extendable/retractable boom thus providing sufficient length so as to have a good effective length at the lower frequencies. During each 15 minute measurement period approximately 160 kilobits of data will be acquired for subsequent transmission to the ground for data processing and analysis.

Directional Spectrum Measurements

Since only a single flight test experiment could be carried out during this phase of the contract, directional spectrum maps of the test area were unable to be generated. This was due to the following causes:

(1) There was sufficient noise on the aircraft 28V. D.C. power buss to render the sweep timing data unusable for controlling the A/D converter.

(2) The presence of this noise component on the receiver IF section electronics casts serious doubts on the validity of the fine structure of the recorded receiver voltage data.

(3) The data were obtained during a period of mild seas which results in a marginal signal-to-noise ratio on the lower frequency channels even if no other problems existed.

(4) There are sufficient uncertainties with respect to the speed and amplifier configuration of the instrumentation recorder available on the aircraft to suggest this as an added source of noise contamination. In addition, the record bandwidth available for this test was insufficient for the data channels from the receiver and sweep controller. It is advisable to have a recorder dedicated to future experiments so that (a) the same recorder can be used for both record and playback and (b) the recorder is available before the flights so that amplifier alignment and bandwidth may be verified.

A plethora of scheduling and operational difficulties prevented the acquisation of more than one data set. If data could have been acquired on two separate flight tests with sufficient time between to correct any problems as originally planned, directional spectrum maps could have been generated.

Recommendations

The difficulties encountered during the flight test can be overcome by the following specific steps:

(a) Isolate the receiving system from the aircraft 28V. D.C. power buss.

(b) Provide a separate portable data recorder dedicated to the bistatic system.

(c) After the above steps, conduct additional flight tests during periods of higher sea state.

Even though these would result in the acquisition of useful data, a number of additional steps are recommended which would simplify the operation of the system, improve the reliability of the data, and allow the system to be routinely operated by personnel other than BCL scientists. These include:

(1) Replacing the Pickard-Burns frequency standard used at the transmitter with a unit having specifications at least as good as the HP-106B used at the receiver.

(2) Replace the R-390 receiver with a solid state IF strip thus eliminating the need for the phase-locked-loops, a potential source of trouble in the aircraft environment.

(3) Simplify the procedure for loading the channel starting frequencies into the digital controllers.

(4) Provide some means for verifying on the aircraft that the correct frequencies are loaded into the receiver controller.

(5) Provide built-in A/D conversion in the receiver and record the data in digital form.

(6) Acquire a more efficient transmitting antenna which could be simply installed at places other than the CPB light tower, for example, a small ship.